Electronic technology

principal contents

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FIVE SHILLINGS

JUNE 1961 VOLUME 38 NUMBER 6

'ENGLISH ELECTRIC'

VAPOUR COOLED POWER TRIODES

STAND No. 1/15

R E C M F

The range of vapour cooled power triodes by the English Electric Valve Company Limited now comprises five standard production types, for r.f. heating and telecommunications equipments, with anode dissipations from 10 to 50 kW.

Vapour cooling, under certain operational conditions, offers positive advantages over other available forms of cooling.

★ Eliminates the need for costly air-filtration in dirt-laden and polluted atmospheres.
★ Effects substantial savings where water costs are high.
★ Can be operated where water is scarce.
★ Ensures completely silent operation.

Abridged data are given here. For full information on these vapour cooled valves and their water and air cooled variants please write to the address below.

E.E.V. Type	Frequency (Mc/s)	Anode Voltage (kV)	Anode Dissipation (kW)	Output Power (kW)
BT189	30	12	35	82
BY194	5	15	30	159
BT1102	20	12	25	53
BY1121	30	10	18	50
BY1122	5	13	10	39

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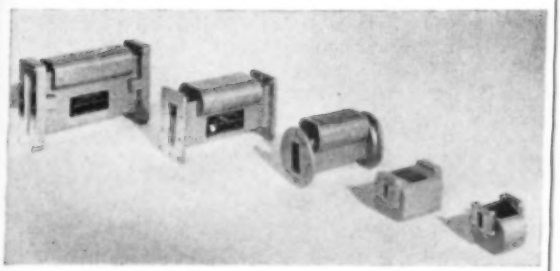

Existing Marconi designs cover most of the frequencies from 1.8 Kmc/s to 10 Kmc/s, and enable the following performance characteristics to be obtained :
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Forward losses of 0.5 dB or less.
A wide range of ferrite devices, including circulators, is available; if your requirements fall outside our existing range we shall be pleased to consider a design to suit you.

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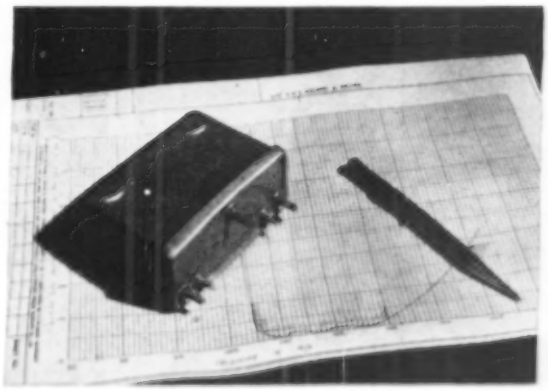
Designed to combine low VSWR with positive mechanical connection at 5000 mc/s, this connector is unique in its use for laboratory measurements and for any application calling for a stable and reliable match. Alternative arrangements are available for flexible cables or panel mounting use. The connectors are normally female but can be converted to male by the simple addition of a locating ring and connecting pin. The design incorporates a sealing ring for waterproofing the connecting joint.

Miniature Crystal Filters



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Marconi's are constantly engaged in the development of new crystal filter techniques, and it is impossible to specify the large number of filter designs available. The Company's unique experience in this field enables us to offer advice and technical assistance with filter problems of any kind.

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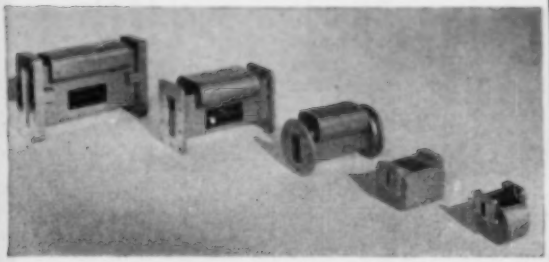

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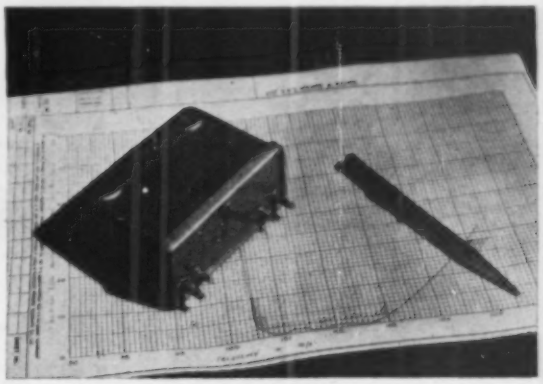

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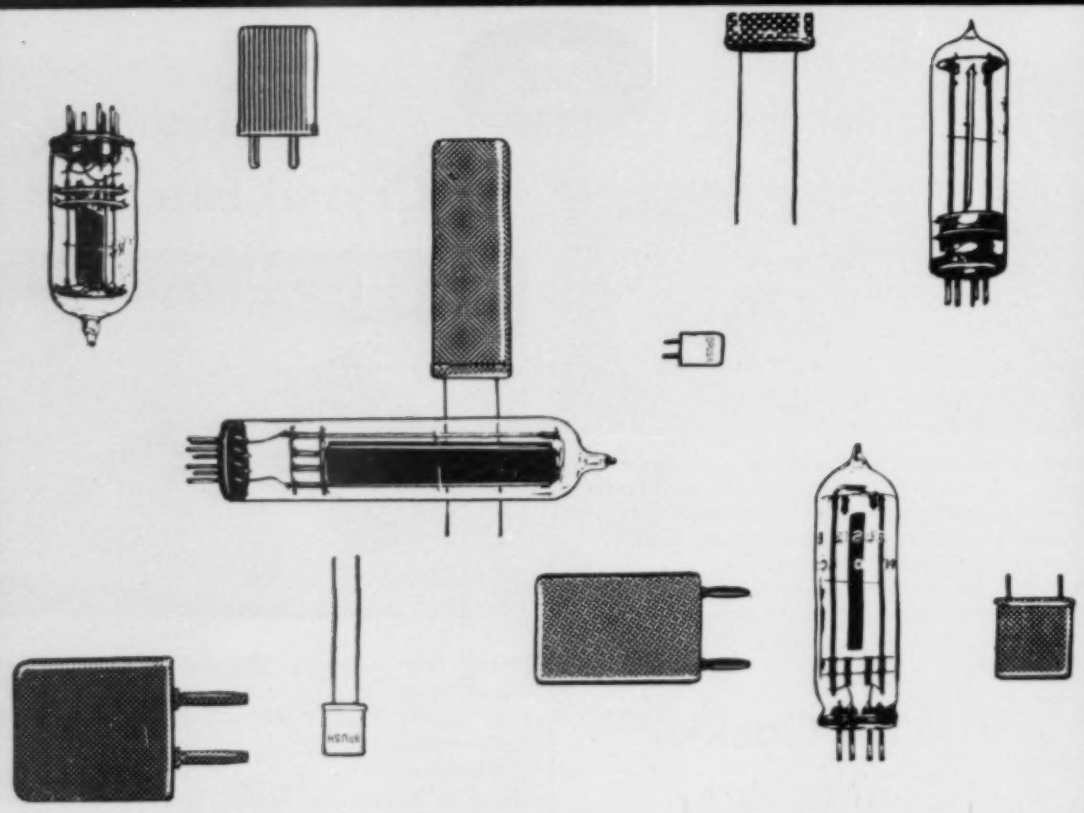

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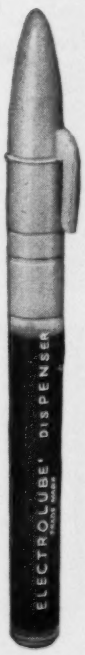

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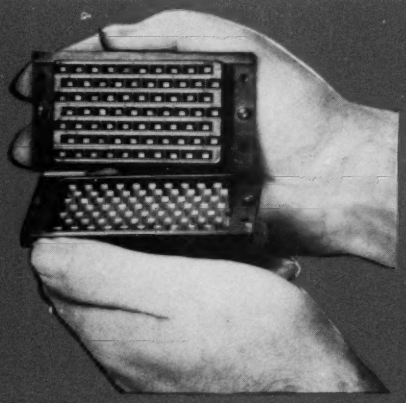

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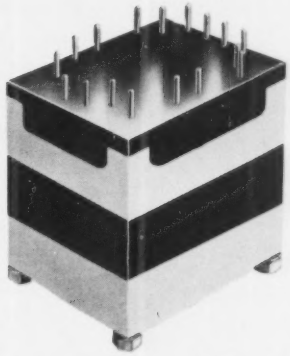

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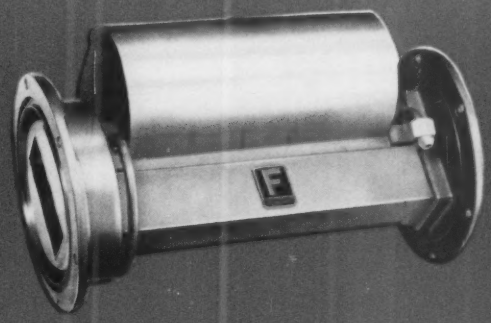

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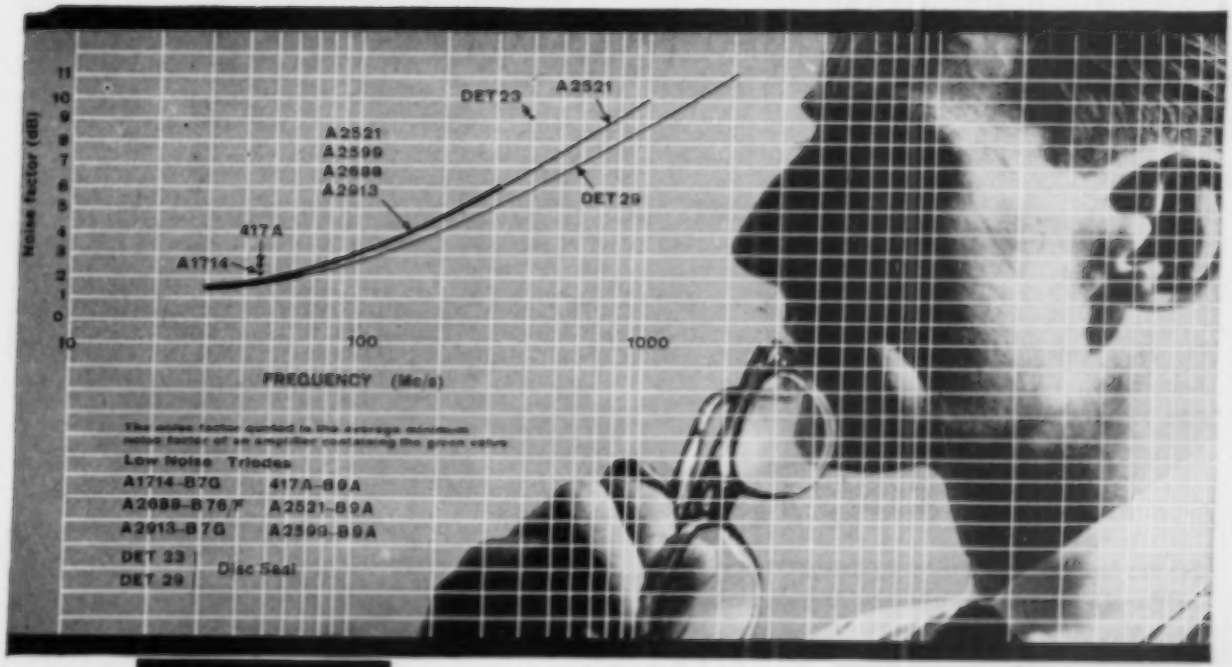

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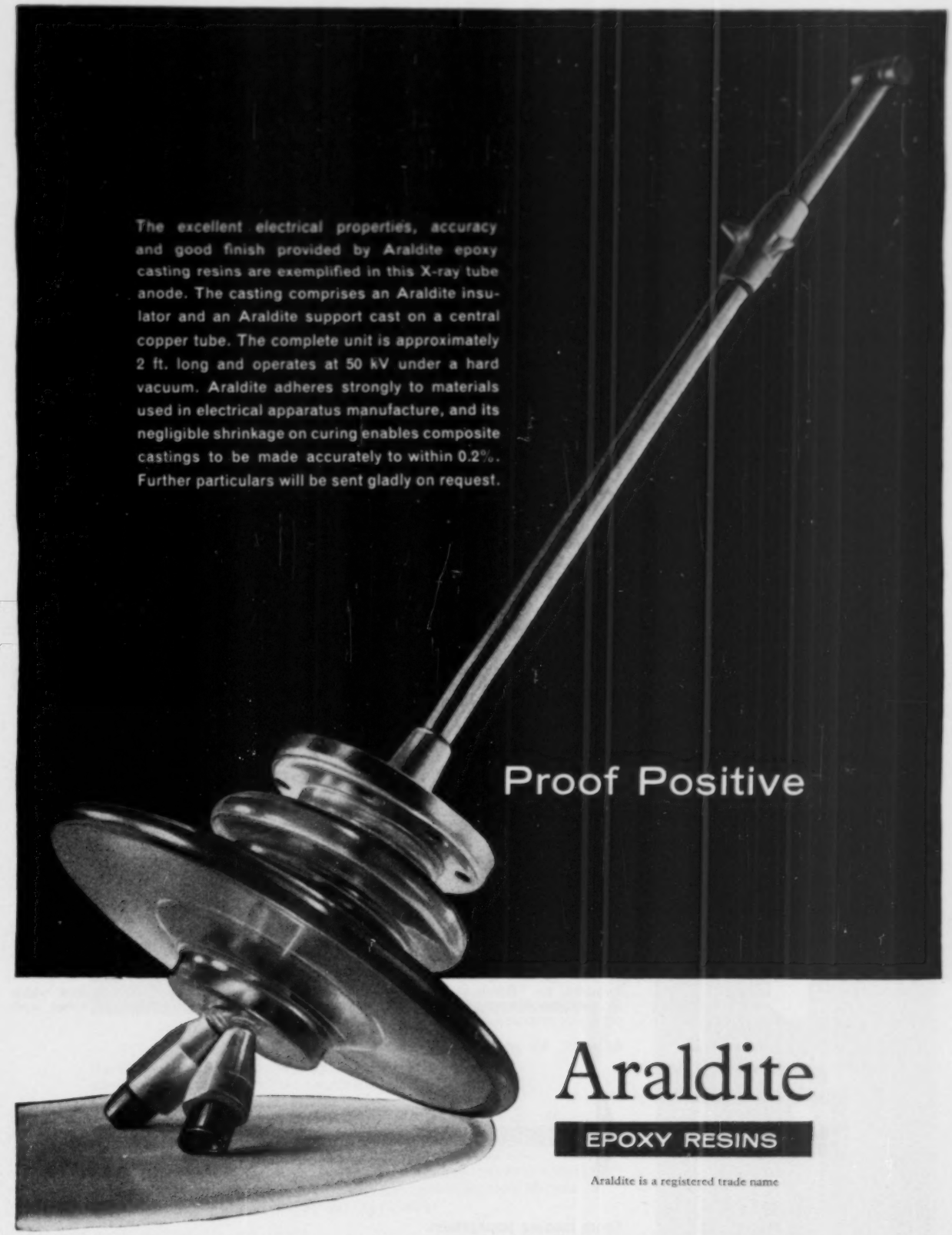

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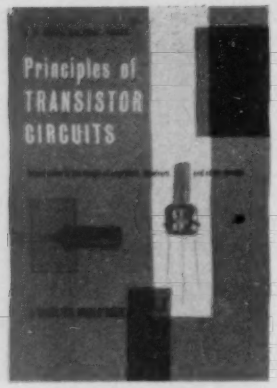

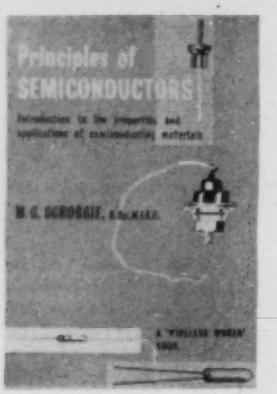

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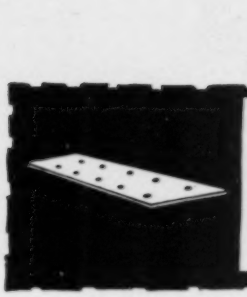

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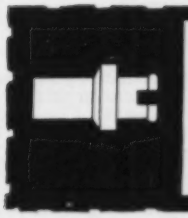

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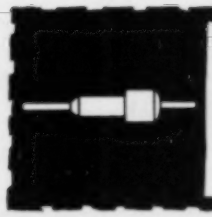

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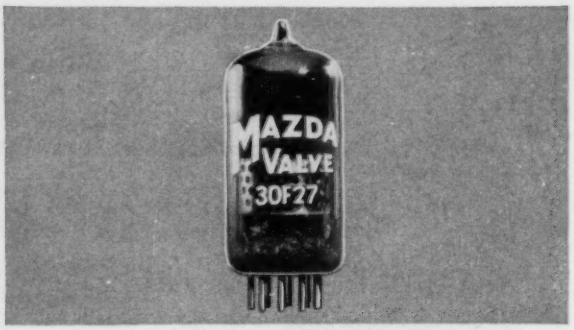
This tetrode used in the RF stage of a television tuner offers certain advantages over the more conventional double triode cascode arrangement. For instance, the number of circuit components required is smaller, the layout is simpler and for valves having comparable slopes the single cathode type valve can be manufactured more economically.

Normally the noise performance of a tetrode is inferior to that of a triode due to the presence of partition noise arising from the screen current. However the 30F27 has been specially designed to provide a low ratio of screen to anode current to minimise the effect of partition noise while still retaining good screening between control grid and anode. When this is used in conjunction with frame-grid techniques a high slope per milliamp of anode current can be obtained resulting in a high gain RF valve with a noise performance much superior to that of a conventional pentode and equal to that of a double triode cascode amplifier such as the 30L1.

Heater Current (amps)	I_h	0·3
Heater Voltage (volts)	V_h	3·7



TENTATIVE RATINGS AND DATA

Maximum Design Centre Ratings

Anode Dissipation (watts)	$p_{a(max)}$	2·5
Screen Dissipation (watts)	$p_{g2(max)}$	0·4
Anode Voltage (volts)	$V_{a(max)}$	250
Screen Voltage (volts)	$V_{g2(max)}$	230
Heater to Cathode Voltage (volts rms)	$V_{h-k(max)rms}$	90*
Cathode Current (mA)	$I_{k(max)}$	18

* From Cathode to higher potential heater pin.

Inter-electrode Capacitances (pF)†

Input Capacitance	c_{in}	6·3
Output Capacitance	c_{out}	1·8
Grid 1 to Anode	c_{g1-a}	0·027
Grid 1 to Grid 2	c_{g1-g2}	2·0
Grid 1 to Cathode	c_{g1-k}	4·0

† Measured in fully shielded socket, without can.

Maximum Dimensions (mm)

Overall Length	56
Seated Height	49
Diameter	22·2

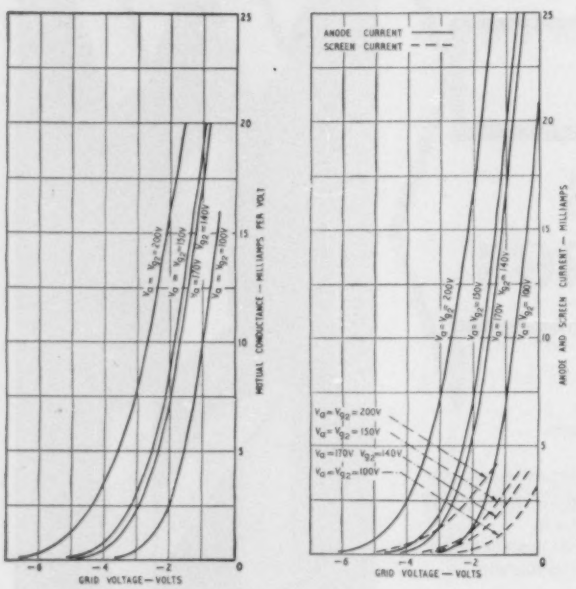
TYPICAL OPERATION

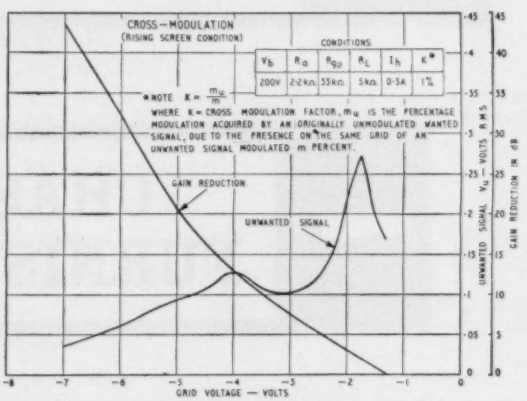
		Cathode Self Bias Circuit	Grid Current Bias Circuit
Supply Voltage (volts)	V_b	200	200
Anode Voltage (volts)	V_a	170	150
Screen Voltage (Initial) (volts)	V_{g2}	140	105
Anode & Screen Common Decoupling Resistor (kΩ)		—	3·3
Anode Decoupling Resistor (kΩ)		2·2	—
Screen Decoupling Resistor (kΩ)		33	33
Cathode Bias Resistor (Ω)	R_k	82	—
Grid Current Bias Resistor (kΩ)	R_{g1}	—	330
Grid Bias Voltage approx. (volts)	V_{g1}	−1·25	—
Anode Current (mA)	I_a	13·5	14
Screen Current (mA)	I_{g2}	1·7	1·4
Mutual Conductance (mA/V)	g_m	15	15·5
Inner Amplification Factor (g_1 to g_2)	μ_{g1-g2}	60	—
Equivalent Grid Noise Resistance (Ω)	R_{eq}	450	—
Input Loss at 50 Mc/s (kΩ)	$r_{g1-k(w)}$	6·8¶	—
Input Capacity Working (pF)	$c_{in(w)}$	10·3¶§	—
Change in Input Capacity produced by biasing valve to cut-off (pF)	$\triangle c_{in(w)}$	2·9¶	—

¶ Measured at 50 Mc/s with the three cathodes strapped and taken directly to earth.

§ Inter-electrode capacity with holder capacity balanced out.

Tentative Characteristic Curves of Ediswan Mazda Valve Type 30F27


MUTUAL CONDUCTANCE — MILLIAMPS PER VOLT
GRID VOLTAGE — VOLTS
ANODE CURRENT
SCREEN CURRENT
ANODE AND SCREEN CURRENT — MILLIAMPS
GRID VOLTAGE — VOLTS



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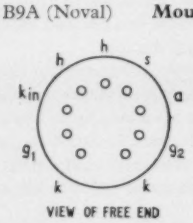

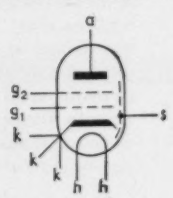

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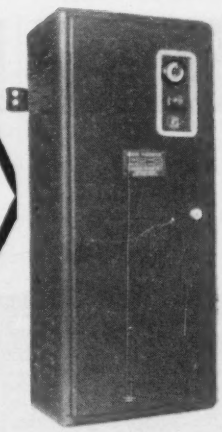

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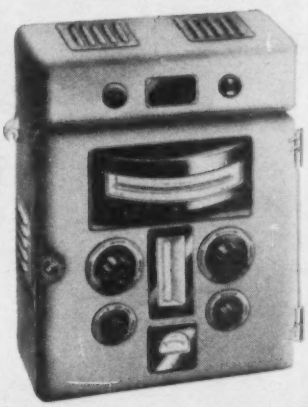

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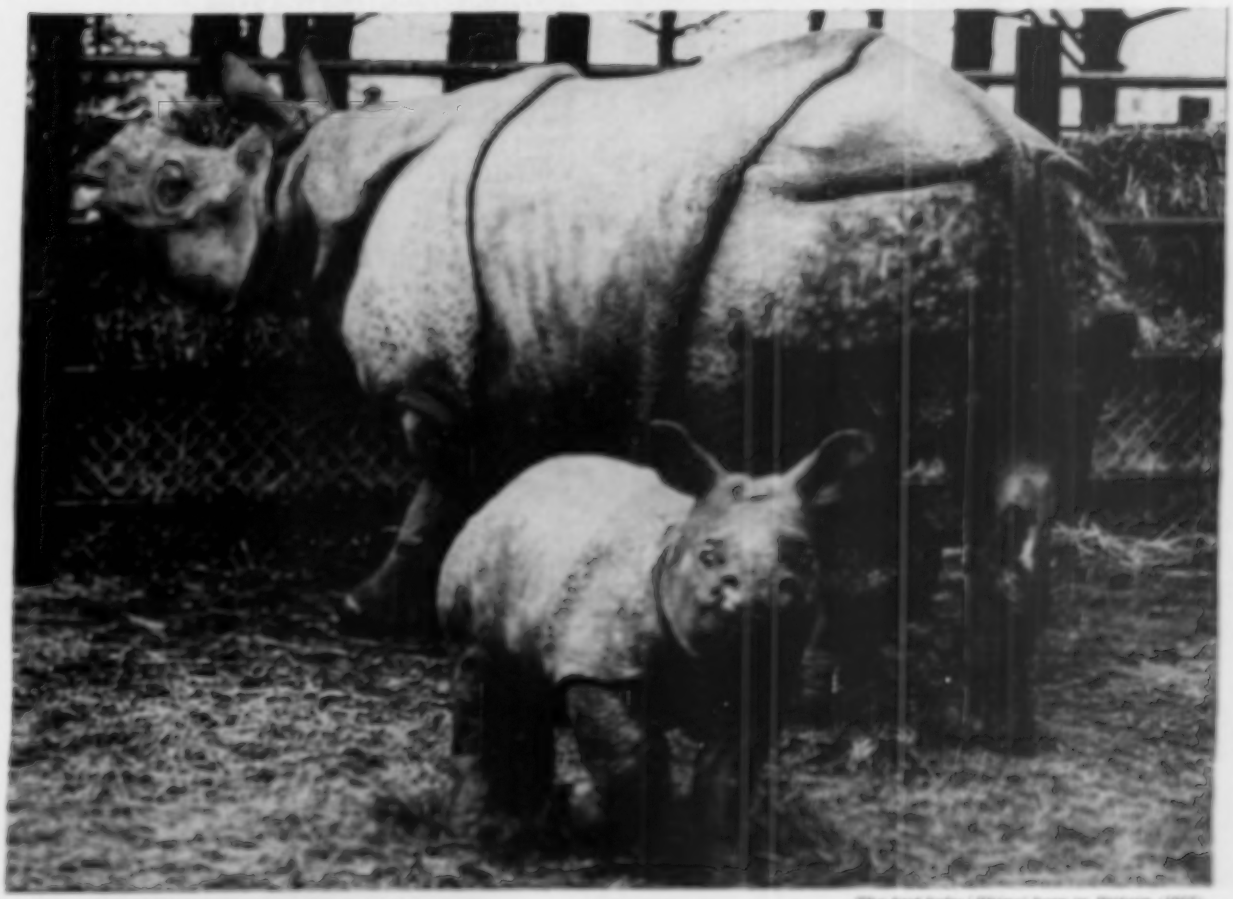

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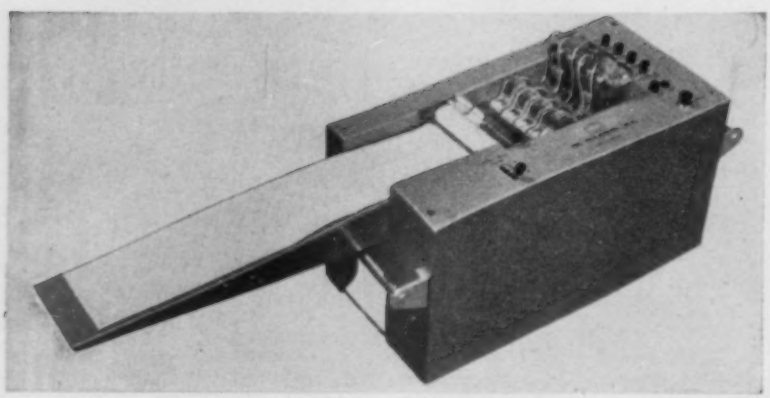

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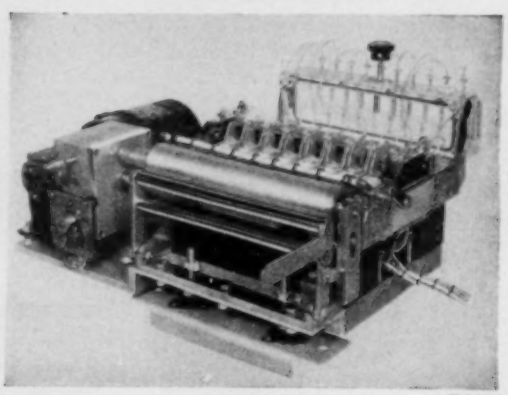

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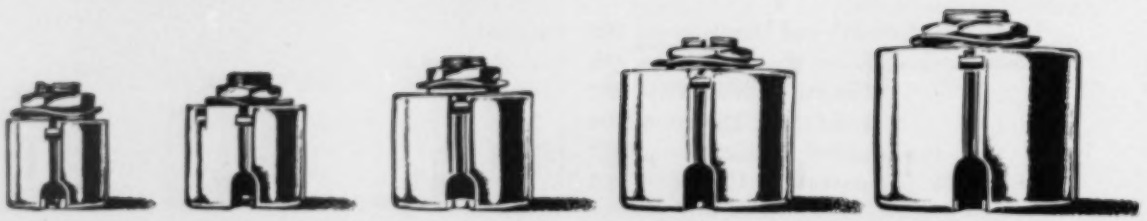

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Electronic technology

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Research and Development

The story of the development of the transistor in the Bell Laboratories in the U.S.A. is well known. This was a brilliant achievement arising from a large co-ordinated effort on the part of physicists, chemists, metallurgists and engineers. It culminated in the appearance of the point-contact transistor in 1948.

Everyone knows this story and a remarkable one it is. Yet how many people know that much of the basic theory of semiconductors was produced in Great Britain? The work of A. H. Wilson on metals and semiconductors was published as long ago as the thirties.

In his book "Electrons and Holes in Semiconductors", Shockley writes in a footnote on p. 10 "The basic theory of semiconductors, which gives analytic form to the ideas discussed here, is due to A. H. Wilson, *Proc. Roy. Soc.*, 1931, Vol. 133A, p. 458". The title of this paper was "The Theory of Electronic Semiconductors".

It is worth enquiring why there is often so long an interval between research and development, for this example is by no means an isolated case. Sometimes, of course, it is because the results of a particular piece of research are in themselves of no practical value but must await the outcome of other work, sometimes in other fields, and it is only the combination of the results which can lead to anything practical.

We suspect that this is not always the case, however, and that development is delayed because of a lack of communication in suitable form of the results of research. The research worker himself does not always have the right outlook to see the practical applications of his work. He is necessarily a specialist, and his paper is usually written for other specialists and is sometimes intelligible only to them.

The people who have practical ideas seem often to be those with a broad outlook and the ability to see relations between widely different things. Such people cannot always understand specialist papers and certainly have not the time to read and comprehend more than a few.

We suggest that it might do much to narrow the gap between research and development if every research worker produced, as well as his specialist paper, a simple version of it written specifically for people who have had no more than a good general scientific training. It should be neither abstract nor summary but an entirely separate paper which avoids specialist jargon and it should be written in words rather than symbols. This may introduce a lack of precision and the account may be incomplete. That does not matter. The purpose of the paper is to inform the general scientific reader so that he may be stimulated by it to produce ideas. When he has an idea he can always turn to the specialist paper for more detailed information and if he is unable to understand it himself he can usually obtain specialist help.

Photoelectric Cells and Photomultipliers

1: VACUUM PHOTOELECTRIC CELLS

By J. Sharpe, B.Sc., A.M.I.E.E.*

It is the aim of this survey to present in simple terms a discussion of the properties and applications of photoelectric cells and multiplier photo-tubes, with especial emphasis on the advantages and limitations of the latter. The treatment is therefore limited to photoemissive devices, in which electrons produced by the action of light are emitted from the sensitive surface into a vacuum (or near vacuum), to be collected by a positive electrode, and the wide field of photoconductive and photovoltaic devices is not considered here.

In order to avoid frequent digressions and footnotes it is proposed first to remind the reader of some aspects of human vision and of the units used in photometry and also to recapitulate some of the elementary physics of quantum theory and of the photoelectric process.

Quantum Nature of Radiation

The fundamental physical phenomenon that determines many aspects of our subject is the fact that electromagnetic radiation such as visible light appears to interact with matter in 'quanta' of fixed energy content, related to the free-space wavelength λ by the equation

$$E = \frac{hc}{\lambda}$$

The visible spectrum extends over the wavelength range 0·4 μ (blue) to 0·76 μ (red), where 1 $\mu = 10^{-4}$ cm. The velocity of light in free space c is 3×10^{10} cm/sec, and h, Planck's constant, is $6{\cdot}55 \times 10^{-27}$ erg-sec, so that a quantum of blue light of wavelength 0·4 μ has an energy of $4{\cdot}9 \times 10^{-12}$ erg. This very small quantity of energy is more conveniently expressed in units such as the 'electron volt', which is the work done in moving an electron through a p.d. of 1 V, and is equal to $1{\cdot}6 \times 10^{-12}$ erg. The quantum of wavelength 0·4 μ is thus of energy 3 eV, while that of 0·8 μ is 1·5 eV. (A convenient formula is E eV $= 1{\cdot}234/\lambda$ where λ is in μ.)

In values of more familiar magnitude, a simple calculation shows that 1 mW of light power corresponds to an energy rate of 6×10^{15} eV per sec, and is carried by 2×10^{15} quanta (or photons) of blue light.

Human Vision

If light of constant power, but of variable wavelength, is allowed to fall on an average human eye, the sensation of brightness which is produced will not be of constant magnitude but varies over the spectrum, being very low or zero for wavelengths longer than 0·76 μ and shorter than 0·4 μ, and a maximum for the region around 0·55 μ as shown in Fig. 1, providing the light is of reasonable intensity[1]. Under these conditions, the central (foveal) region of the eye is employed. At very low light levels, receptors situated in the peripheral regions of the eye (rods) take over, the peak of the sensitivity curve shifts to 0·510 μ, and the perception of colour differences is much reduced. (This is the Purkinje effect, which accounts for the distinctive appearance of a landscape by moonlight).

The energy needed to stimulate the sensation of brightness is about 100 times smaller in this latter case than with foveal vision, and at 0·510 μ it is found that flashes of light giving an energy of between 2 and 6×10^{-10} erg at the pupil of the eye are visible with a probability of 50%. Measurements of this kind are made with apparatus providing light flashes at a level which is reduced in steps until, over a number of observations at constant intensity, half of the flashes are just visible.

The fact that this statistical type of result is obtained is a direct consequence of the quantized nature of light interaction and is due to the small number of quanta actually involved in the stimulation of sensation. Since

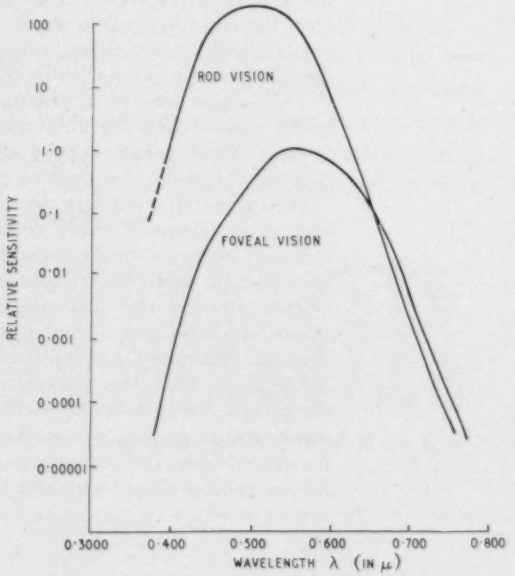

Fig. 1. *Variation of eye sensitivity versus wavelength (average of many observers) for foveal and rod vision*

* *E.M.I. Electronics Ltd.*

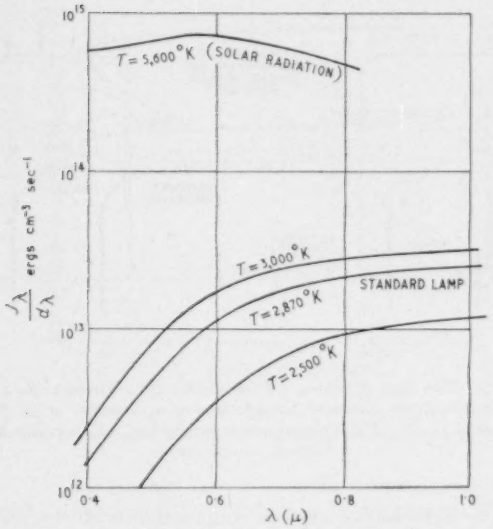

Fig. 2. *Spectral distribution of power emitted from black bodies at various temperatures*

one photon at 0·510 μ is equivalent to $3{\cdot}9 \times 10^{-12}$ erg, the light flashes referred to above contained between 50 and 150 photons at the surface of the eye, and it can be shown that the number actually effective in the sensitive material of the retina lies between 5 and 15 photons[2].

Since each of these is absorbed independently of any other in the flash, the number absorbed will show a statistical variation from flash to flash which can be described by Poisson's formula $P_x = n^x/(\epsilon^n\, x!)$. Here, P_x is the probability of x photons being absorbed, when the average number absorbed is n. If n is the minimum number of photons required to produce visual sensation, and for a particular observer is equal to 6, then the total probability of absorbing less than 6 quanta is

$$\sum_0^5 P_x = \sum_0^5 6^x/(\epsilon^6\, x!) \simeq 0{\cdot}5$$

so that there is a 50 : 50 chance of observing the flash, which is in accordance with experiment.

(The above statistical concepts will prove of use in dealing with photoelectron emission in later sections.)

Photometry

In most visual tasks the light levels are well above the limiting values, however, and the normal units of photometry involve light powers of the order of milliwatts.

The original, pre-electrical, basis of visual photometry was the 'candle', which is now defined in terms of standard electric lamps. A source of one candlepower which emits light uniformly in all directions is said to produce a luminous power of 4π lumens (it will be seen that this is effectively a reciprocal definition of 'candle' and 'lumen'), and for light of wavelength 0·550 μ, the lumen corresponds to a power of 1·6 mW. For light of other spectral composition, a greater radiant power is necessary to give an equivalent visual sensation and, for instance, one lumen of sunlight contains 4 mW. The upper curve in Fig. 2 shows the approximate spectrum of sunlight, and corresponds to that emitted by a black body at a temperature of 5600 °K. The lower curves are for black-body radiation at colour temperatures to be found in tungsten lamps, the value of 2870 °K being that used in standard lamps. (The physical temperatures would be higher than the 'colour temperatures' quoted, which are those of radiating black bodies having the appropriate colour.)

As the colour temperature decreases, the proportion of power emitted at the red end of the spectrum increases, and over the wavelength range of 0·4 to 0.76 μ, one lumen from a 2870 °K radiator corresponds to 6 mW while, at 2500 °K, 8 mW is needed. It will be seen that in spite of the higher radiant power in the latter case, that available in the blue part of the spectrum, from 0·4 to 0·5 μ, is 20% lower than from a 2870 °K radiator. This factor becomes of importance when photoelectric devices having a spectral sensitivity different from the eye are used for photometry.

Photoelectricity

The absorption of a quantum of electromagnetic radiation at the surface of a suitably-prepared metal or alloy may give rise to the ejection of an electron if the energy hc/λ is greater than some threshold value. It can be seen intuitively that the 'quantum efficiency' of this process (i.e., the number of electrons emitted per incident photon) is likely to be much less than unity, if only because the energy transfer must take place in a layer comparable with λ (i.e., several atoms thick) from which the electrons must emerge and in which scattering away from the surface may occur.

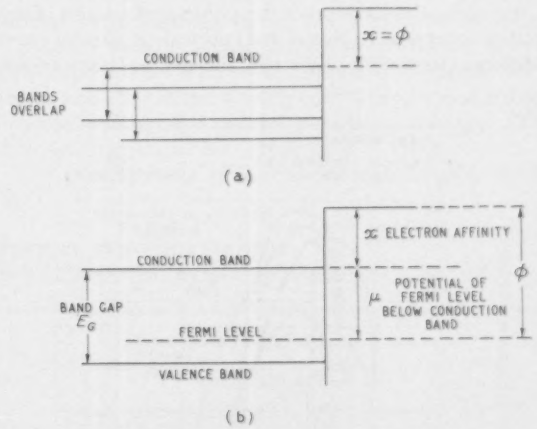

Fig. 3. *(a) Shows the energy levels of metal, such as gold, and (b) shows the energy levels of semiconductors*

It has, in fact, been shown that electrons produced at depth greater than $\simeq 0{\cdot}025\ \mu$ below the vacuum–photocathode interface will not emerge from the surface[3].

The potential energy of electrons at the interface between a solid and vacuum is illustrated in Fig. 3, and this shows that electrons, in the so-called conduction band, which are able to move through the material under the influence of an applied p.d., must overcome a potential barrier before they can emerge into the

vacuum. Other electrons are more tightly bound in a lower level and the gap between the two bands represents the energy which must be supplied to move an electron into the upper (conduction) band. In a normal metal, such as gold, the conduction band is well supplied with electrons, and the value of ϕ, the work function, equal to the electron affinity, is 4·5 eV, so the long wavelength threshold is about 0·27 μ, which is well in the ultra-violet region of the spectrum. At room temperature, the average thermal energy available to any electron in the metal is 0·025 eV so the number of thermionic electrons able to escape through the potential barrier is very small, and is about 1 per minute per cm².

In a semiconductor, such as an alloy of antimony and caesium, Cs_3Sb, photons must provide sufficient energy to move an electron from the Fermi level across the energy gap, and leave it with a surplus to overcome the surface forces, if it is to be emitted as a photoelectron. In Cs_3Sb, the electron affinity is between 0·4 and 0·6 eV, and the energy gap is about 1·4 eV, so that the long wavelength threshold corresponds to an energy around 1·9 eV; i.e., 0·660 μ. The fact that the work function is lower than that of gold makes it easier for some of the free electrons to escape from the surface and, at room temperature, the thermionic emission rate may be between 10^4 and 10^2 per cm per sec, depending on the way the surface has been produced.

Above the photoelectric threshold, the quantum efficiency for the photoelectric process, that is to say, the ratio of photoelectrons to incident photons, varies with photon energy in a manner, which although complex in detail, follows the general form shown in Fig. 4, rising to a maximum as the excess energy provided to the electron increases, and falling off in a manner determined by the optical absorption of the photo-surface and of any window interposed. Since this spectral-sensitivity curve is different from that of the eye, the spectral composition

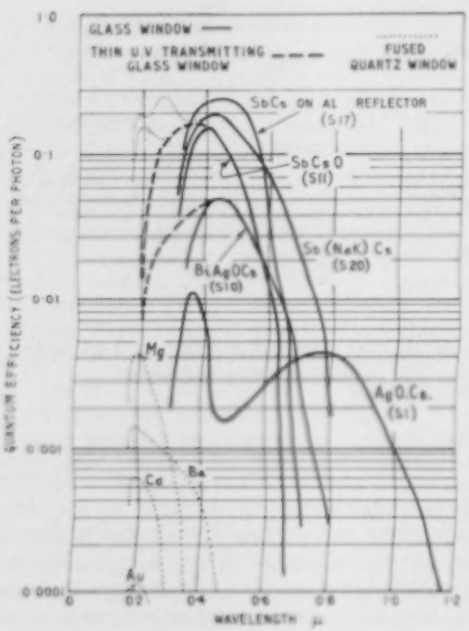

Fig. 4. *Spectral response of various photocathodes*

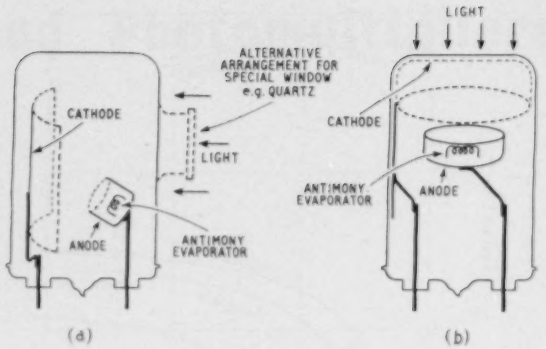

Fig. 5. *Two ways of forming a photocathode. For convenience, the anodes can incorporate an evaporator for deposition of a component of the photo-cathode; e.g., Sb.* [*(a) Opaque-cathode construction, (b) semi-transparent cathode construction*]

of any light source used in conjunction with the photo-emissive material must be carefully considered, particularly if photometric measurements in terms of lumens are to be made. The existence of sensitivity well outside the visual range should also be noted.

Photocells

We are now in a position to consider the parameters of photocells in terms of the phenomena described above. It will be seen that the long-wavelength threshold, the quantum efficiency and its variation with wavelength and the magnitude of any thermionic emission are factors of importance, and to these we must add the technical features of construction.

Two ways of forming a photocathode are available and are shown in Fig. 5 (a) and (b). The opaque cathode (a) is deposited on a metal electrode, so that the electrons must emerge from the side on which the light is incident. The semi-transparent form (b) is deposited on a transparent substrate, which is usually the window through which light must pass. In this case, the electrons emerge into the vacuum from the surface opposite to that receiving the radiation. A modification of this arrangement is one in which a semi-transparent cathode is deposited on a highly reflecting layer of evaporated aluminium to give a more efficient arrangement than the normal opaque layer.

An anode must be mounted in the envelope in order to collect the photoelectrons, and typical arrangements are shown in the diagram.

The current–voltage characteristics of typical vacuum photo-diodes are shown in Fig. 6, for various values of incident illumination. At a sufficiently high value of anode voltage, all the photoelectrons are collected but, at lower voltages, the field near the cathode is insufficient to overcome negative space charge. It is therefore important to design a photocell so that the collecting field is reasonably uniform over the cathode surface so that the lowest saturation voltage can be obtained.

At high cathode currents, the resistance of the photo-surface may be sufficiently great to give a deviation from linearity, and this is particularly true of semi-transparent surfaces, so that there is a limiting luminous flux appropriate to each type of cathode surface.

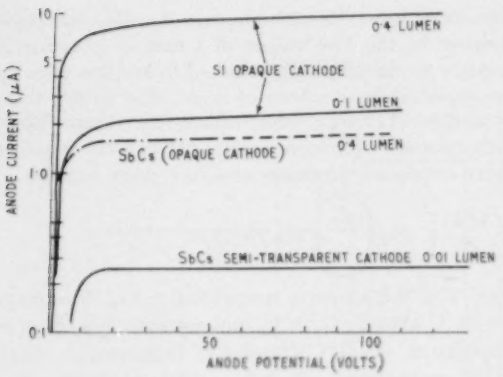

Fig. 6. *Saturation characteristics of typical vacuum photocells*

In the absence of illumination, a small current flows between anode and cathode ("dark current"), which is partly the result of thermionic emission from the cathode, but much more due to leakage currents within the tube, and across the exterior of the envelope, and so can be reduced by internal guard-rings from values around 10^{-9} to less than 10^{-12} A at room temperature.

Various types of photocathode available, together with their spectral response curves are given in Table 1 and Fig. 4, which also show the influence of the window cut-off in the ultra-violet. Photosensitivity values, in terms of microamps per lumen of 2870 °K tungsten light, are tabulated for typical surfaces and the influence of extended red sensitivity in increasing this latter figure should be noted.

It will be seen that no photoemissive surface with sensitivity above 1·4 μ is available, and that the silver-oxide-caesium alloy owes its not unreasonable μA/lumen sensitivity to the fact that tungsten light of 2870 °K colour temperature has a power maximum at around 1 μ. At the other end of the spectrum, normal glass envelopes cut off radiation at around 0·35 μ; special u.v. transmitting glasses cut off at about 0·22 μ, while fused-silica windows transmit down to 0·165 μ, so that we have a wavelength range from 0·165 to 1·4 μ over which photoemissive surfaces in commercially-obtainable photo-cells are of use. Specially made cells with LiF windows may be obtained to reduce the cut-off wavelength to 0·12 μ.

For objective photometry in the visual range, it is possible to correct the spectral response of surfaces such as Bi-AgO-Cs, and Sb-Na-K-Cs, which have adequate red response, to approximately that of the average eye, by appropriate combination of filters. Fig. 7 defines a suitable sector filter[4] for the former surface, and it will be obvious that if this is to be mounted close to an end-window photocathode, of the type shown in Fig. 5 (b) both the distribution of energy in the light beam and the sensitivity over the cathode must be uniform, if accurate correction is to be achieved. It must be borne in mind that the latter condition is rarely obtained, since the techniques for producing most photosensitive surfaces involve the evaporation of a material such as antimony or silver over a quite large area of substrate, which may itself have been subject to processing, and the subsequent alloying of this evaporated layer with one or more alkali metals from the vapour phase. It is therefore not surprising that variations in sensitivity of up to 30% are to be expected and must be allowed for in the design of the optical system used with a photocell. In general, the radiant energy should be defocused, or diffused, to cover as large a sensitive area as possible and, if possible, should not be allowed to move about the cathode.

Measurement of Photocurrent

Under normal conditions, the values of photocurrent to be measured and perhaps recorded will lie between 6×10^{-6} and 10^{-12} A, although special applications, such as those met with in astronomy, may involve currents as low as 10^{-16} A.

The rates of emission of electrons from the photo-surface will therefore range from 1 000 to 10^{13} per sec, and the statistical nature of the phenomenon sets a limit to the accuracy of measurement, particularly in the lower ranges; it is this aspect of measurement with which this section is particularly concerned.

The idealized circuit of Fig. 8(a) shows a photocell connected in series with a capacitance C, and a polarizing battery. The potential across C is measured by means of a device such as a quartz-fibre electrometer[5] of very high impedance. The switch across C is opened for a time t, and the charge flowing into C produces a voltage V (small compared with the polarizing voltage). The photocell current is then given by $i = VC/t$.

The photocurrent produced by illumination of the

TABLE 1
TYPICAL PHOTOCATHODES
(Glass windows except where noted†)

Photocathode	Form	Peak Sensitivity		Long Wavelength Threshold	Typical Sensitivity (2870 °K lamp)
		λ	Quantum Efficiency		
		μ	Electrons/photon	μ	μA/l.
SbCsO (S–11) ..	Semi transparent	0·42	0·15	0·67	60
SbCs (S–4)	Opaque	0·45	0·10	0·7	40
SbCs (EMI "S") ..	Semi-transparent	0·42	0·12	0·65	40
Sb(NaK)Cs (S–20) ..	,,	0·44	0·20	0·85	150
BiAgOCs (S–10) ..	,,	0·45	0·05	0·8	35
AgOCs (S–1) ..	,,	0·8 (also in blue)	0·004	1·2	15
†Mg (quartz window)	,,	<0·2	0·004	0·39	—
†Au (quartz window)	,,	<0·2	0·0001 to 0·00001	0·27	—

photocathode may then be measured by successive observations of the difference in charge collected on the capacitance C with the cell in the dark, and exposed to illumination.

Under constant conditions, however, it will be found that the observed values of V vary from reading to reading, although over a sufficient number of observations, a definite average value V_0 is obtained. This fluctuation is due to the fact that the charge VC is the sum of a finite number $n_0 = V_0 C/e$, on the average, of individual electrons, which are emitted independently of each other.

As will be seen from the earlier Section (Human Vision), the probability of x electrons being emitted is given by the Poisson formula, and for the relatively large values of n_0 concerned this may be approximated by the Gaussian form

$$P_x = \frac{1}{\sigma\sqrt{2\pi}} \exp. \frac{-(x - n_0)^2}{2\sigma^2}$$

The symbol σ is known as the standard deviation of the distribution and is the root mean square of the deviations of individual readings x from the mean value n_0 and so is a measure of the 'noise' produced by the statistics. For values of n_0 greater than 20, $\sigma^2 = n_0$, so for a current i, measured over a time t, $\sigma = \left(\frac{it}{e}\right)^{\frac{1}{2}}$, and the r.m.s. values of voltage fluctuation $\sigma e/C$ will equal $(eit)^{\frac{1}{2}}/C$, while the average voltage $V_0 = it/C$. The relative r.m.s. error of an individual measurement $\frac{\sigma e}{V_0 C} = \left(\frac{e}{it}\right)^{\frac{1}{2}}$ and so is reduced by integrating the charge over as long time as possible.

Where the dark current i_d of the cell is large enough to make difference methods necessary, the fluctuation in $V - V_d$ will have a standard deviation given by the square root of the sum of the squares of the standard deviations of the two conditions. In one case the current is i_d and, in the other, is $i + i_d$ and the r.m.s. fluctuation Δi is $\left(\frac{i_d e}{t_d} + \frac{(i_d + i)e}{t}\right)^{\frac{1}{2}}$ where t_d and t are the appropriate times of integration.

There may arise occasions when a fixed time T is available for a difference measurement and the current $i + i_d$ is known to be approximately r times the dark current. In this case, it can be shown that the best accuracy will be obtained when the ratio T/t_d is made equal to $(r + 1)^{\frac{1}{2}} + 1$.

Although the drift-rate method gives the highest sensitivity for the measurement of small currents, it is very often inconvenient, particularly when the current is itself changing with time. In this case, the circuit of Fig. 8(b) is used, and the p.d. produced across a resistance R is measured by an instrument having the required frequency response. This may be a quartz-fibre electrometer, or perhaps an electrometer tube, feeding a meter or an oscilloscope with sufficient amplification. In any of these cases, there may be assigned an appropriate bandwidth Δf which is related to the sampling time used in the drift method by the relationship $t = 1/2\Delta f$, and the root mean square fluctuation of the voltage Ri will be $R(2ei\Delta f)^{\frac{1}{2}}$. This represents the noise voltage due to the fluctuations in the rate of emission of electrons from the cathode, the so-called 'shot-noise'.

Owing to the low values of i met in practice, it is necessary to use very high values of R, and it is then found that an additional source of noise, due to the thermal fluctuations of charge density along the resistor, becomes of importance (Johnson noise). It can be shown that a pure resistance produces an r.m.s. noise voltage

$$\left(\frac{4kT}{R}\int_{f_1}^{f_2} R^2 df\right)^{\frac{1}{2}} = (4kTR\Delta f)^{\frac{1}{2}} \quad .. \quad .. \quad (1)$$

Here, T is the absolute temperature (i.e., the temperature in °C above −273 °C, and equal to 300 °K at room temperature of 27 °C), and k is Boltzmann's constant, 1·3708 erg/degree, or $8{\cdot}6 \times 10^{-5}$ electron-volt per degree. At $T = 300$ °K, $kT = 0{\cdot}025$ eV, so that $kT/e = 0{\cdot}025$ V.

The r.m.s. value of Johnson noise then equals $\left(\frac{4kT}{C} eR\Delta f\right)^{\frac{1}{2}} = (0{\cdot}1eR\Delta f)^{\frac{1}{2}}$ V at $T = 300$ °K and the total r.m.s. noise

$$\begin{aligned}(\overline{V_n^2})^{\frac{1}{2}} &= (R^2 2ei\Delta f + 0{\cdot}1eR\Delta f)^{\frac{1}{2}} \\ &= (eR\Delta f)^{\frac{1}{2}}\,(2Ri + 0{\cdot}1)^{\frac{1}{2}} \text{ V} \quad .. \quad .. \quad (2)\end{aligned}$$

from which it follows that Ri must exceed 0·05 V, if Johnson noise is not to be predominant.

If a thermionic electrometer tube is used, another source of noise will be present, due to the fluctuations in the rate of gain and loss of charge on to the control grid. Even if the net grid current is zero, this will be due to the fact that positive charges are arriving at the grid at the same average rate as negative charges, and the sum of these two components will make up the effective grid current i_g, which may be as low as 2×10^{-15} A in an electrometer tube, or as high as 10^{-8} A in a conventional tube. The shot noise due to the grid current now adds a component $(R^2 2ei_g\Delta f)^{\frac{1}{2}}$, making the total noise voltage

$$(\overline{V_n^2})^{\frac{1}{2}} = (CR\Delta f)^{\frac{1}{2}}\,[2R(i + i_g) + 0{\cdot}1]^{\frac{1}{2}} \quad .. \quad (3)$$

It will be seen that for many purposes i_g may be ignored, providing a suitable electrometer tube is available.

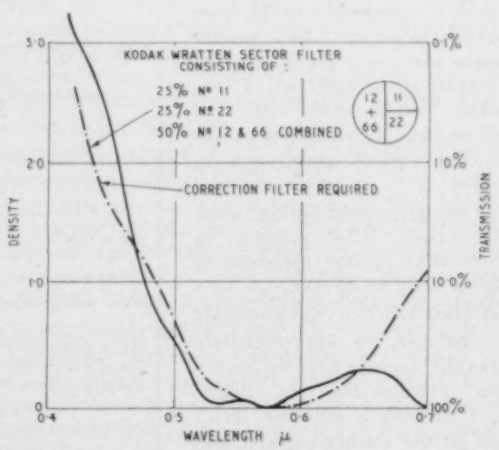

Fig. 7. *Filter for correction of BiAgOCS (S10). End window photocell response to standard eye*

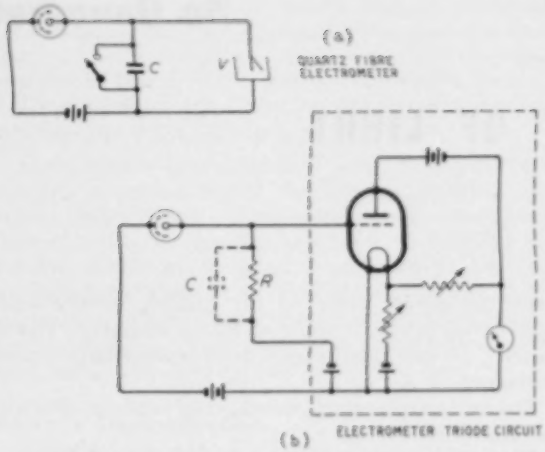

Fig. 8. *Simple circuits for the measurement of photocurrent; (a) rate of drift, (b) continuous current*

In practical circuits, the load resistor R will always be shunted by a capacitance C, made up from the capacitances of the circuit components and leads, and the value of R^2 in the integral of Equ. (1) must be replaced by the real part of the impedance $|Z|^2$ so that

$$(\overline{V_n^2})^{\frac{1}{2}} = (2ei + 4kT/R)^{\frac{1}{2}} \left(\int_{f_1}^{f_2} |Z|^2 df\right)^{\frac{1}{2}} \quad .. \quad .. \quad (4)$$

For a parallel RC circuit $|Z|^2 = R^2/(1 + 4\pi^2 f^2 R^2 C^2)$; the integral $(f_1 = 0, f_2 = \infty)$, becomes $R/4C$, and the r.m.s. noise voltage is

$$(\overline{V_n^2})^{\frac{1}{2}} = \left(\frac{eR}{4RC}\right)^{\frac{1}{2}} (2Ri + 0{\cdot}1)^{\frac{1}{2}} \quad .. \quad .. \quad (5)$$

If Equ. (5) is compared with Equ. (2), it will be seen that $1/4RC$ has replaced $\Delta f (= f_2 - f_1)$ and, in fact, if undue frequency distortion is to be avoided, $2\pi f_2 RC$ must be less than unity.

Since the response time t of the measuring instrument is set by RC, we may put $R = t/C$, and then the minimum value of current i_{mtn} to give noise comparable with the Johnson noise is $i_{mtn} = C/(20t)$ and the noise voltage is approximately $90\,C^{-\frac{1}{2}}$ V if C is in farads.

For a fairly typical C of 25 pF, $(\overline{V_n^2})^{\frac{1}{2}} = 18\ \mu$V and i_{mtn} is 10^{-12} A for $t = 1$ sec, and 1 μA for $t = 1\ \mu$sec.

By the use of negative feedback[6] involving an amplifier of loop gain A, the time constant at the input may be reduced by a factor of $1/A$ (providing $A \gg 1$) so, for a given response time t, the resistance R may be increased by a factor A, and i_{mtn} correspondingly decreased to a value at which the limit is set by grid current i_g. Under these circumstances, the best results may be obtained by the use of a vibrating-reed electrometer[7] in which the effects of grid current are eliminated.

The results of the above discussion may usefully be applied to a few examples to illustrate the limiting conditions: (a) An intensity of $1{\cdot}5 \times 10^{-10}$ lumen, equivalent to the light received from the Polestar (or a candle at 6,000 ft.) by a 1-in. telescope, is to be measured as accurately as possible, at intervals of 60 sec. A cell with a sensitivity of 60 μA/L and a dark current of 2×10^{-15} A is available, so the signal current is 9×10^{-15} A and, r, the ratio of $i + i_d$ to i_d is 1·45. The dark current is then integrated for a time $60/[1 + (2{\cdot}45)^{\frac{1}{2}}] = 24$ sec, and the signal plus dark current for 36 sec. The uncertainty in a single difference reading is then $1{\cdot}6 \times 10^{-17}$ A, or about 0·2%. If an integrating capacity of 25 pF is used, the voltages recorded are 19·3 and 41·9 mV respectively.

For a star of magnitude, giving an intensity of 10^{-12} lumen, the signal current would be 6×10^{-17} A, and r is then so close to unity that the time is divided equally between the two measurements and the uncertainty is $(4i_d e/T)^{\frac{1}{2}}$, or $1{\cdot}5 \times 10^{-17}$ A, or 25% of the signal. In this case, to obtain an accuracy of 1%, T should be increased by a factor 25^2 and, in place of 1 min, we have over 10 hr.

(*b*) A fluorimeter utilizing a cell of dark current 10^{-12} A and sensitivity 30 μA/L, is to be used with a simple electrometer circuit of input capacitance 25 pF and i_g 10^{-14} A. The meter response time is to be 1 sec. It is required to estimate the amount of meter fluctuation for various light levels, and to compare the result with that given by using a feedback circuit of loop gain 100. In the simple circuit $RC = 1$ sec, so $R = 4 \times 10^{10}\ \Omega$, and the r.m.s. noise voltage,

$$(\overline{V_n^2})^{\frac{1}{2}} = \left(\frac{1{\cdot}6 \times 10^{-19} \times 4 \times 10^{10}}{4 \times 1}\right)^{\frac{1}{2}} [2 \times 4 \times 10^{10}$$

$$(i + 10^{-14}) + 0{\cdot}1]^{\frac{1}{2}} = 4 \times 10^5\,(8 \times 10^{10}\,i + 0{\cdot}1008)^{\frac{1}{2}}$$

For $i = 0$, we have a noise voltage of 12·5 μV, while for $i = 2{\cdot}5 \times 10^{-13}$ A, we have 14 μV. The voltage Ri for the latter case is 10 mV, and if this is made to be full scale deflection, the fluctuation is 0·14% of full scale, the latter corresponding to a light intensity of 10^{-8} lumen. At one-tenth full scale, the fluctuation is 1·3% of the reading.

For the feedback circuit, R may be increased by a factor 100, and the noise voltage is $4 \times 10^{-4}\,(8 \times 10^{12}\,i + 0{\cdot}18)^{\frac{1}{2}}$ μV giving 160 and 580 μV at zero light and full scale, respectively. Ri is now however 1 V at 10^{-8} lumen, and the fluctuation is 0·06% at full scale, and 0·2% of the reading at 10^{-9} lumen.

(c) A simple flying-spot television system utilizing a photocell with a straightforward amplifier of bandwidth 1 Mc/s is required to give a signal (power)-to-noise (power) ratio of 1,000 or better at peak white. The constants of the system are required.

Since RC is about 5×10^6 sec^{-1}, R is about $10^4\ \Omega$ for $C = 25$ pF. Assuming that valve noise can be neglected, we have a noise voltage of $(2 \times 10^{-9})^{\frac{1}{2}}$ $(2 \times 10^4\,i + 0{\cdot}1)^{\frac{1}{2}}$, so that for the required signal-to-noise ratio,

$$\frac{(10^4 i)^2}{2 \times 10^{-9}\,(2 \times 10^4 i + 0{\cdot}1)} = 1{,}000,$$

so that i for peak-white is approximately 0·05 μA, giving an input voltage of $\frac{1}{2}$ mV. The r.m.s. voltage is almost entirely determined by the Johnson noise in the resistor and is about 15 μV, so the picture would be very noisy in the blacks.

(*To be concluded*)

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THE VELOCITY OF LIGHT

Even today you can still find people who will tell you that physics is a dull subject. Lack-lustre is certainly in the eye of the beholder, and it may possibly stem from a certain confusion between the simple language of science and the dreadful jargon that the copywriters persuade people to accept as having some meaning. The Principle of Inertia, and the Conservation of Energy; what more shining prospects could a man want? Give yourself a book of Harnwell beneath the bough . . .; and it is disappointing to find that it isn't quite like that. The Exclusion Principle and the Principle of Uncertainty —master these, and you can vacillate the *easy* way; but there is no easy way when you get beyond merely dithering. The Second Law of Thermodynamics; well, what have Clausius and Boltzmann got that Parkinson hasn't got? Relativity seems a write-off; what is the use of learning how to keep up with the Joneses when the Joneses are bursting their seams learning to keep up with you? But wait. There is, after all, your own *special* Relativity, for the finer spirit, the urbanely discriminating man who insists on the best of everything; what a let-down when you find that it leads straight to a formula for mass production!

Again, it may be that these ingrates feel that, as things are discovered and filed away, they become part of an ever more complicated dogma which is to be accepted without question. The excitement of the chase is all for the hunter; the student sees only the stuffed bodies in their glass cases. This, of course, is quite a wrong approach, and as the whole of twentieth-century science shows, the carefully embalmed principles of the nineteenth century have one by one got up and slowly walked away. Will this, in due course, happen to the lively captures of more recent times? We do not know, and can only suspect that it very well may do.

Where is all this leading? Well, in March of this year, a hand-out from the Institute of Radio Engineers suggested that we may have to revise our ideas on the subject of Special Relativity and the constancy of the velocity of light. There are some points here that I do not quite understand, so that it seems best to quote it in full to avoid misrepresentation.

"Unsuspected Variation in Jupiter Eclipses Raises Question about Einstein Theory on Velocity of Light

New York, N.Y., March 22, 1961—A twenty-two year series of observations of the eclipses of the satellites of Jupiter, made over a half century ago, may contradict Einstein's Special Theory of Relativity.

According to Martin Ruderfer, president of Dimensions, Inc., of Brooklyn, N.Y., in a technical paper presented today to the annual International Convention of the Institute of Radio Engineers, recent analysis of the observations show unsuspected timing variations.

Future interplanetary travel could well be affected, the author stated.

The observations, which were made by the Harvard Astronomical Observatory between 1878 and 1900, were originally analyzed by Ralph Sampson, a British astronomer, and published in 1909. Sampson's results were subjected to an additional analysis by Mr. Ruderfer, who found unsuspected variations of up to 1½ seconds in the timing of the eclipses that depended upon the positions of Jupiter and the earth in their orbits.

Mr. Ruderfer's analysis was undertaken to determine if the velocity of light is constant in all directions of space as required by Einstein's Special Theory of Relativity.

Previous measurements of the velocity of light have been made with a two-way light signal which travels out-and-back. Although this average velocity has always been found to be constant, it is not known for certain that the separate out and back portions each have identical speeds.

Astronomical observations should permit a one-way measurement of the velocity of light and therefore a new fundamental check on relativity.

If a variation in the velocity of light is found, it would support the validity of a rival theory, the Fitzgerald-Lorentz contraction theory.

The author stated that the observed timing variations allow the possibility that there is a variation in the velocity of light for different directions in space. But, it was pointed out, if such an explanation is correct it cannot completely explain the observed timing variations.

A partial explanation may lie in the inaccuracy in the distances of Jupiter and the earth to the sun. Another contributing factor may be seasonal variations on the surface or in the atmosphere of Jupiter.

According to Mr. Ruderfer, a more complete analysis with a computer is required to precisely explain and to distinguish between the possible causes of the timing variations. Besides allowing a new kind of check on the theory of relativity, it would provide a more accurate estimate of the basic Astronomical Unit of distance.

The Astronomical Unit is equal to the mean distance of the earth from the sun—approximately 93 million miles—and is a basic yardstick for measuring distances within the solar system.

Exact knowledge of this distance is important for interplanetary travel. Considerable information pertain-

ing to Jupiter's weather should also be obtained.

In addition, there is the possibility that some information on the geographical features of the four large satellites of Jupiter is contained in the eclipse data.

These satellites are about the size of our moon and are therefore potential space stations."

There seem to be a number of points to take up here. Römer's determination of the velocity of light; this "rival theory", the "Fitzgerald-Lorentz contraction theory"; how far the special theory of relativity has hitherto stood up to attack; and finally, the real possibility that there really may be asymmetries in the propagation of light that may open new fields of physics for investigation. Need I say that I don't know very much about any of these except the first two.

Römer's Determination

This is described in all the elementary textbooks, and the original communication appeared in the *Journal des Scavans*, December 1676. I have been reading a reproduction in I. Bernard Cohen's book "Roemer and the First Determination of the Velocity of Light" (Burndy Library Inc., New York, 1944), which gives a number of Römer's actual observations, and also mentions their contemporary reception. You can see the general idea from Fig. 1. The first satellite goes round Jupiter, J, in roughly the same plane as the eliptic in about 42½ hours; during this time the earth moves from A to B or, six months later, from E to F. The eclipses occur at shorter intervals on the AB side than they do on the EF side, because of the finite time taken by light to cover the distance BA or EF. The cumulative effect, over a large number of eclipses, amounts to an unpunctuality of about 16½ minutes if you start counting at H and calculate the time of eclipse to be observed six months later at G. So that it takes light some 16½ minutes to cover the diameter of the earth's orbit. Römer's actual figure was about 22 minutes at first; Newton, in the 1704 edition of the "Opticks", would make it about 14 to 16 minutes. When all the variable quantities are taken into account, it is not surprising

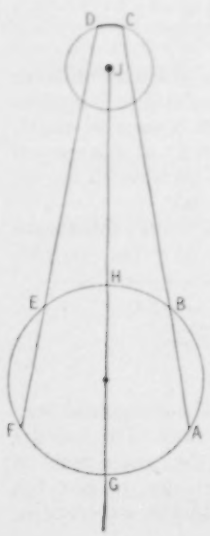

Fig. 1. *Jupiter J with a satellite in relation to the earth's orbit A B E F G. This is Römer's original picture; the centre of the big circle is, of course, the sun*

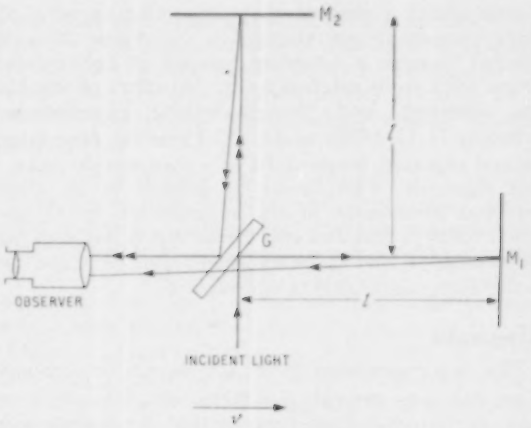

Fig. 2. *Scheme of the Michelson-Morley experiment. Here G is a beam-splitter, and after traversing their separate paths GM_1G parallel to the direction of the earth's orbital motion and GM_2G at right angles to it (or in any case, two paths at right angles to one another), the two beams are superposed to give an interference pattern in the field of observation*

that the astronomers of the time viewed the result with very great reserve; and it was not until 1729, when Bradley obtained an independent value for the speed of light from the aberration constant, that they all really firmly accepted Römer's "delay of light" as being attributable to the finite speed of light in space.

The Michelson-Morley Experiment

To refresh your memory, Fig. 2 shows the scheme of the Michelson-Morley experiment. The path from G to M_1 and back is in the direction of v, the earth's orbital velocity of about 30 km.sec^{-1} round the sun. The effect sought for, assuming that v and c, the velocity of light, were measured with respect to a fixed ether, was an interference-fringe displacement due to the difference in time for the two double journeys. For GM_1G, the time is

$$t_1 = \frac{l}{c-v} + \frac{l}{c+v} = \frac{2l}{c\left(1 - \frac{v^2}{c^2}\right)};$$

while, with the above assumption, the velocity of light over the path GM_2G relative to the apparatus is $\sqrt{c^2 - v^2}$, and the time is

$$t_2 = \frac{2l}{\sqrt{c^2 - v^2}} = \frac{2l}{c\sqrt{1 - v^2/c^2}}$$

Expanding the expressions for t_1 and t_2, and going only as far as the first-order terms for v^2/c^2, the time difference $(t_1 - t_2)$ is $\frac{l}{c} \cdot \frac{v^2}{c^2}$, which with the apparatus used by Michelson and Morley should have given a path difference of about half a wavelength. Rotation of the apparatus through 90°, so that M_1 and M_2 interchange roles, should make this path difference the other way round while, with GM_1 and GM_2 both at 45° to v, the path difference should vanish. Without going further into things, it can be seen that as the system is

rotated about a vertical axis perpendicular to v, the fringe pattern should change in the course of a 90° rotation through a complete reversal of light-for-dark fringes back to its original form. No effect of this kind was observed, and though several experimenters, including D. C. Miller and C. V. Drysdale, have sought for and reported fringe shifts of a comparable order to that expected, there seems at present to be almost universal acceptance of the experiments by G. Joos which confirm that any effect observed is less than one-thousandth of that to be expected. But we come now to the Fitzgerald-Lorentz contraction.

Fitzgerald

The two expressions for t_1 and t_2 can be reconciled if we can only get rid of a factor of $\sqrt{1 - v^2/c^2}$, and this can be done if we suppose that the path-length l which is moving with velocity v has shrunk to $l\sqrt{1 - v^2/c^2}$ as a result of motion. The outcome of this is that, whether there is a fixed ether with reference to which absolute motion can be recorded, we should never be able to measure or detect this anyhow, since the measuring-rod changes its length suitably so as to frustrate this. Really, Fitzgerald considered the contraction to result from a change in the intermolecular forces within the material of the rod, so that there was a distortion of shape that might be expected to depend on the properties of the material, and be different for different substances. Rayleigh showed, for example, that transparent materials should develop an anisotropy that would reveal itself as a detectable double refraction; but none of these expectations was fulfilled, and this one of Rayleigh's required some further looking into. Eventually, Larmor in 1904 suggested that "the additional hypothesis that matter is electrical, that is, containing electrons" would reconcile things. Which is just about where we came in, a beautiful balance of the equation $0 = 0$. Lorentz will appear in a moment when we have nailed down the really crucial point about the Fitzgerald contraction. It had to invoke a sort of misbehaviour of *materials* because it refused to take liberties with *time*. The Fitzgerald contraction is wedded to absolute time.

Now, Einstein did not discover relativity or relative motion. Newton was very clear that the laws of motion that he propounded were laws of relative motion. But Newton was perfectly happy about absolute time, which simply flowed uniformly. Einstein realized that both simultaneity and the lengths of observed time intervals were different in different frames of reference. Accepting this, the *relativity of time* (which is the essential twentieth-century idea), the appropriate calculations to reconcile observations of the same event in different frames of reference are given by the Lorentz-transformation formula.

The Lorentz Transformation

Nothing was said in the previous paragraph about the constancy of the velocity of light in all directions. It was, of course, assumed—presumably from symmetry, or because there seemed no reason otherwise. But the next stage in the argument, the Lorentz transformation, assumes that the speed of light is not only the same in all directions, but is the same *whatever the frame of reference*

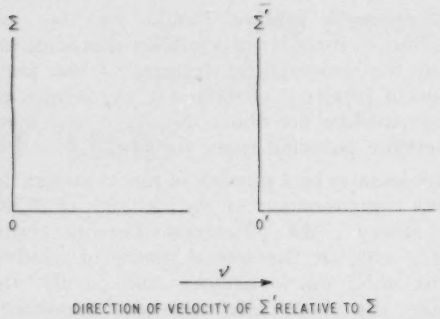

Fig. 3. *The frame of reference Σ′, in which an observer makes observations x′ and t′, moving with a velocity v relative to the frame Σ, in which the observations are x and t*

within which the observations are made. If, as Fig. 3, for example, we have two frames Σ and Σ', such that Σ' is moving with a uniform velocity v relative to Σ in the x-direction (this is not cheating by taking a particular case, since there can only be motion in one direction at once, and we make this the x-direction) then if $dx/dt = \pm c = dx'/dt'$, we have $x^2 - c^2t^2 \equiv x'^2 - c^2t'^2 = 0$ for the two frames. The condition for this identity to hold is that x, x' and t_1, t' are related by the transformations

$$x' = (x - vt)/\sqrt{(1 - v^2/c^2)}$$

and

$$t' = (t - x\,v/c^2)/\sqrt{(1 - v^2/c^2)}.$$

Now, the Lorentz contraction does not really suppose that anything changes in shape or size as the result of its motion. In fact, whether this could happen or not is outside its scope. What it says is that the simultaneous observation of the two ends x_1, x_2 of a rod in Σ of length $l = (x_2 - x_1)$ will not give the same result as the simultaneous observation of the same two ends, now x_1' and x_2', in Σ', which give $l' = (x_2' - x_1')$; and that the reason is that the two systems interpret simultaneity differently because of the relativity of time.

The argument goes on as follows:
From the equation for x',

$$x_2' - x_1' = \frac{(x_2 - x_1) - v(t_2 - t_1)}{\sqrt{(1 - v^2/c^2)}}$$

Here, t_1 and t_2 are the instants in Σ at which an observer in Σ' sees the measuring-rule with one end at 0 applied to x_1 and x_2 respectively. But what is seen as simultaneous from Σ' is not simultaneous in Σ; so it is not just a question of putting $t_1 = t_2$; what we have to do, for simultaneity in Σ', is to make $t_1' = t_2'$.

Now, from the equation for t', if $t_1' = t_2'$, this means $t_1 - x_1\,v/c^2 = t_2 - x_2\,v/c^2$, so that $(t_2 - t_1) = (x_2 - x_1)v/c^2$, whence

$$x_2' - x_1' = \frac{(x_2 - x_1)\,(1 - v^2/c^2)}{\sqrt{1 - v^2/c^2}}$$

so

$$l' = l\sqrt{(1 - v^2/c^2)}$$

Now you can see the difficulty. The Fitzgerald contraction is a construction in absolute time. The Lorentz contraction is a construction in the space-time of relativity. They do happen, numerically, to give the same answer that fits the Michelson-Morley experiment.

I find it hard to see how any Fitzgerald-Lorentz contraction can conflict with or "rival" special relativity. But it is easy to get confused, and get not only the wrong contraction but the wrong Fitzgerald; for there was, you remember, the one whose observations on the first satellite of Jupiter must have conflicted with Römer's—he spoke of a moon of his delight that hath no wane.

The Constancy of *c*

But the points raised in the I.R.E. hand-out are serious ones. Dr. C. V. Drysdale spent a good deal of time and effort towards the end of his life in just such a search, for a means of shaking the special theory of relativity. Professor G. Temple ("Turning Points in Physics", North-Holland Publishing Co., 1958, p. 76) said: "Much as I should like to disprove the special theory of relativity, and in spite of the many years I have given to this task, I have to admit that I cannot find any flaw in the evidence. I really think it is true that there is a genuine null effect, and that predictions based on the Maxwellian electrodynamics and an absolute frame of reference are in fact erroneous". M. E. J. Gheury de Bray (famed about fifty years ago as the author of "Exponentials Made Easy" which, together with Silvanus Thompson's "Calculus Made Easy", must have helped many scientists to take their first steps in mathematics) believed in about 1930 that he had discovered a secular decrease of about 4 km/sec per year in the value of c, which would have light brought to rest in a mere 75,000 years from now. But a critical examination of all the data then available by R. T. Birge in 1941 showed that this was simply a secular decrease in the errors of observation which usually tended to overestimate.

But there is another line of approach. By c we mean the velocity of light in vacuo and, presumably, also in field-free space. Quantum electrodynamics predicts a possible slowing-down of light in a transverse magnetic field, with several processes which might be used to detect this. Professor T. Erber of Chicago has discussed some of these possibilities briefly ("Velocity of Light in a Magnetic Field", *Nature*, 1st April 1961, Vol. 190, p. 25) and it might be worth while to examine some of his conclusions later on. Here, it is sufficient to say that he is dealing in terms of rotations of the plane of polarization of the order of 10^{-6} to 10^{-7} radians, and spectroscopic resolving powers of the order $\Delta\nu/\nu = 10^{-16}$. When the departures from the "constancy of c" that quantum electrodynamics predict are themselves so small that techniques which could detect them have still to be developed, it doesn't seem that the satellites of Jupiter are a very promising field to explore. But you never know. And in any case, nobody can seriously say that physics is a dull and unenterprising study when page one of the elementary light textbooks springs vigorously to life in such a way.

NEW H.F. DEPLETION-LAYER TRANSDUCER

Bell Telephone Laboratories have announced that they are developing a high-frequency transducer which, at present, will operate efficiently at about 830 Mc/s, but the frequency range is expected to be extended to above 10,000 Mc/s. This is a piezoelectric device utilizing a semiconductor depletion layer. It is anticipated that the primary application of this transducer will be for ultrasonic delay lines where its operation at high frequencies and wide bandwidths will make it possible to store large amounts of information.

The transducer consists of a plate of piezoelectric semiconductor material (such as gallium arsenide) on which a thin metal film is deposited (Fig. 1). The film constitutes a non-ohmic rectifying contact which causes a depletion layer to form. The thickness of the depletion layer can be controlled with a negative-bias voltage across the interface.

When an alternating voltage is applied, most of the voltage drop occurs across the layer and the layer behaves in a manner similar to a very thin piezoelectric crystal which is bonded to a solid. Since the layer is thin, 10^{-3} to 10^{-5} cm, the electric field is very large and considerable piezoelectric stress can be produced in the layer.

It is claimed that the depletion-layer transducer has several advantages over an ordinary transducer—viz., (a) because the layer is so thin the greatest efficiency of the transducer is at very high frequencies (this may be as much as 100 times more efficient than any other known transducer when operating at 10,000 Mc/s); (b) the voltage-dependent thickness of the layer, hence the resonant frequency of the transducer can be varied by varying the d.c. bias voltage; (c) present early-development models measured at 600 Mc/s have a bandwidth of 5% (30 Mc/s), an order of magnitude larger than typical ceramic transducers, operating at frequencies below 10 Mc/s. This means that a comparable increase should be expected in the amount of information which can be transmitted and, finally, (d) the transducer is relatively simple to make and easy to handle.

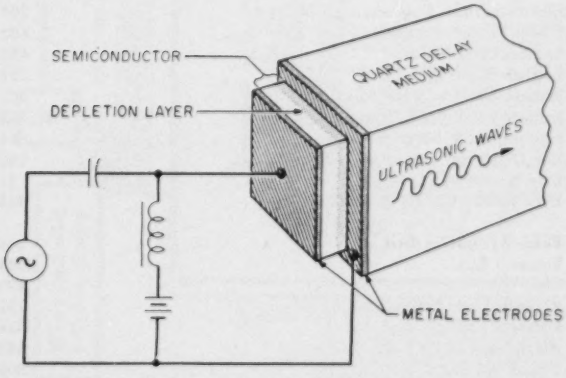

Fig. 1. *A schematic diagram of the Bell Telephone depletion-layer transducer attached to a quartz crystal*

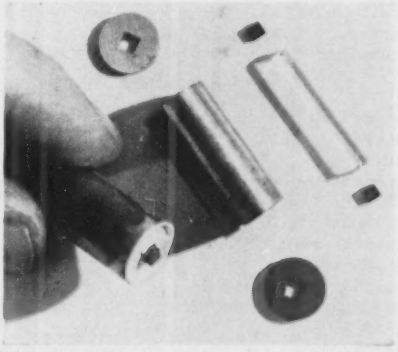

Fig. 2. *An exploded view of an ultrasonic delay line using two depletion-layer transducers; a one-inch long quartz crystal serves as the delay medium*

R.E.C.M.F. Show 1961

AT OLYMPIA, LONDON, W.14, FROM 30th MAY TO 2nd JUNE

Hours of opening, 10 a.m. to 6 p.m. (closes at 9 p.m. 31st May)

THE R.E.C.M.F. Exhibition, promoted by the Radio and Electronic Manufacturers' Federation, is being held at Olympia this year. Admission is either by invitation from the exhibitors or by payment of 5s. at the door for U.K. visitors; overseas visitors receive a season ticket free of charge.

This issue of *Electronic Technology* contains a list of exhibitors and a plan showing the layout of the exhibition. The following pages include illustrations and descriptions of a few of the items being shown. This year, as in every past year, the latest developments of the industry are being displayed but, undoubtedly, the accent is on reliability and the use of new materials and techniques to achieve it. In addition, because of the increased display area, many new instruments and methods are being demonstrated which have been designed to increase the speed and reliability of component testing.

LIST OF EXHIBITORS

EXHIBITORS	STAND NO.
A.B. Metal Products Ltd.	210
A.K. Fans Ltd.	353
Air Control Installations Ltd.	358
Aircraft-Marine Products (G.B.) Ltd.	301
Alma Components Ltd.	478
Amphenol (Great Britain) Ltd.	226
Ancillary Developments Ltd.	476
Andec Ltd.	455
Anderton Springs Ltd.	351
Antiference Ltd.	354
Ardente Acoustic Laboratories Ltd.	100
Ariel Pressings Ltd.	404
Arrell Electrical Accessories Ltd.	207
Arrow Electric Switches Ltd.	500
Arrow Ltd.	498
Ashburton Resistance Company	473
Associated Electrical Industries Ltd.	162
Associated Iliffe Press Ltd.	489
Autop Foundry	458
Aveley Electric Ltd.	360
Avo Ltd.	225
BICC-Burndy Ltd.	156
Bader Machinery Co. Ltd.	494
Bakelite Ltd.	491
Beckman Instruments Ltd.	471
Belclere Co. Ltd., The	472
Belling & Lee Ltd.	313
Berec International Ltd.	401
Beulah Electronics	152
Bird, Sydney S., and Sons Ltd.	109
Birmingham Sound Reproducers Ltd.	161
Bradley, G. & E., Ltd.	456
Bray, Geo & Co. Ltd.	108
Brayhead Electronics Ltd.	318
Bribond Ltd.	57
Brimar Ltd.	274
British Electric Resistance Co. Ltd.	103
British Insulated Callender's Cables Ltd.	156
British Physical Laboratories	479
Brush Crystal Co. Ltd.	317
Bulgin, A. F., & Co. Ltd.	315
Burndept Ltd.	52
C.C.L. Ltd.	457
Carr Fastener Co. Ltd.	202
Cathodeon Crystals Ltd.	362
Chapman Ultrasonics Ltd.	152
Ciba (A.R.L.) Ltd.	228
Clarke, H., & Co. (Manchester) Ltd.	157
Colvern Ltd.	154
Connollys (Blackley) Ltd.	171
Continental Connector Ltd.	402
Cosmocord Ltd.	213
Dagnall & Kendall Ltd.	102
Darwins Ltd.	409
Datum Metal Products Ltd.	450
Dawe Instruments Ltd.	212
Day, J., & Co. (Derby Works) Ltd.	110
Department of Scientific & Industrial Research	462
"Diamond H" Switches Ltd.	254
Direct T/V Replacements Ltd.	152
Doran Instrument Co. Ltd.	152
Dubilier Condenser Co. (1925) Ltd.	214
Duratube & Wire Ltd.	252
E.M.I. Sales & Service Ltd.	105
Egen Electric Ltd.	265
Ekco Plastics Ltd.	271
Electrical Remote Control Co. Ltd.	411
Electro Acoustic Industries Ltd.	259
Electrolube Ltd.	474
Electro Methods Ltd.	223
Electronic Components	488
Electronic Technology	489
Electrothermal Engineering Ltd.	204
Elliott Brothers (London) Ltd.	482
Enalon Plastics Ltd.	406
English Electric Co. Ltd.	221
English Electric Valve Co. Ltd.	165
Enthoven, H. J., & Sons Ltd.	268
Equipment & Services Ltd.	411
Erg Industrial Corporation Ltd.	256
Erie Resistor Ltd.	54
Ever Ready Co. (G.B.) Ltd.	401
Fane Acoustics Ltd.	364
Ferranti Ltd.	366
Fine Wires Ltd.	270
Flotorm Parts Ltd.	256
Formica Ltd.	206
Fortiphone Ltd.	352
Fox, F.X., Ltd.	350

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Electronic technology
First Floor
Ground Floor
Addison Road
RESTAURANT
MAIN ENTRANCE
GRILL
POST OFFICE
COCKTAIL LOUNGE
LADIES
GENTLEMEN
EXHIBITORS CLUB

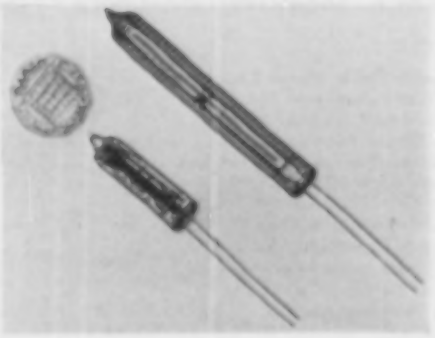
R.E.C.M.F. Show 1961 Continued

EXHIBITORS	STAND NO.
Gardners Radio Ltd.	475
Garrard Engineering & Manufacturing Co. Ltd.	275
Gayler & Hall Ltd.	453
General Controls Ltd.	173
General Electric Co. Ltd.	163
Goldring Manufacturing Co. (G.B.) Ltd.	208
Goodmans Industries Ltd.	264
Gramophone Co. Ltd.	105
Greencoat Industries Ltd.	111
Gresham Transformers Ltd.	460
Guest, Keen & Nettlefolds (Midlands) Ltd.	251
Haddon Transformers Ltd.	276
Hallam, Sleigh & Cheston Ltd.	224
Harwin Engineers Ltd.	492
Hassett & Harper Ltd.	408
Hellermann Ltd.	269
Henry & Thomas Ltd.	357
Herbert Publishing Co. Ltd.	114
Heywood & Co. Ltd.	107
Hinchley Engineering	164
Hughes International (U.K.) Ltd.	499
Hunt, A. H. (Capacitors) Ltd.	166
Imhof, Alfred, Ltd.	159
Imperial Chemical Industries Ltd., Plastics Division	309
Insulating Components & Materials Ltd.	151
International Rectifier Co. (G.B.) Ltd.	158
Ionic Plating Co. Ltd.	251
Irish Cables Ltd.	229
Jackson Brothers (London) Ltd.	263
Jobling, James A., & Co. Ltd.	220
K.L.G. Sparking Plugs Ltd.	458
L.S.B. Components Ltd.	152
Labgear Ltd.	362
Langley London Ltd.	464
Lewis Spring Co. Ltd.	113
Linton & Hirst Ltd.	150
Linvar Ltd.	469
Livingston Laboratories Ltd.	451
London Electric Wire Co. & Smiths Ltd.	262
London Electrical Mfg. Co. Ltd.	311
Lucas, Joseph, (Electrical) Ltd.	456
Lustraphone Ltd.	272
M.C.P. Electronics Ltd.	468
M-O Valve Co. Ltd., The	163
M.S.S. Recording Co. Ltd.	477
McMurdo Instrument Co. Ltd.	400
Magnetic Devices Ltd.	279
Magnetic & Electrical Alloys Ltd.	306
Mallory Batteries Ltd.	222
Mansol (Great Britain) Ltd.	104
Marconi Instruments Ltd.	314
Marrison & Catherall Ltd.	466
Measuring Instruments (Pullin) Ltd.	55
Mec-Test Ltd.	486
Metway Electrical Industries Ltd.	470
Microcell Electronics	481
Microvernier Ltd.	224
Midland Silicones Ltd.	459
Miniature Electronic Components Ltd.	486
Ministry of Aviation	461
Minnesota Mining & Mfg. Ltd.	169
Morganite Resistors Ltd.	305
Mullard Ltd.	160
Mullard Equipment Ltd.	491a
Multicore Solders Ltd.	308
Multimusic Ltd.	308
Murex Ltd.	206
Mycalex & T.I.M. Ltd.	253

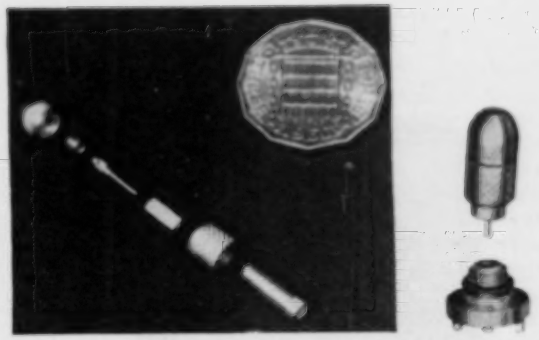
EXHIBITORS	STAND NO.
N.S.F. Ltd.	219
Neill, James, & Company (Sheffield) Ltd.	227
Newmarket Transistor Co. Ltd.	361
Oliver Pell Control Ltd.	483
Painton & Co. Ltd.	155
Parmeko Ltd.	168
Permanoid Ltd.	207
Plannair Ltd.	53
Plessey Co. Ltd.	310
Preformations Ltd.	310
Pressac Ltd.	410
Printed Circuits Ltd.	495
Pye Switch Division	362
Redcliffe Industries (Bristol) Ltd.	483
Rejafix Ltd.	359
Reliance Cords & Cables Ltd.	312
Reliance Manufacturing Co. (Southwark) Ltd.	273
Rendar Instruments Ltd.	363
Reproducers & Amplifiers Ltd.	403
Reslosound Ltd.	152
Richard Allan Radio Ltd.	112
Rivlin Instruments Ltd.	407
Rola Celestion Ltd.	255
Ross, Courtenay & Co. Ltd.	493
Salford Electrical Instruments Ltd.	216
Salter, George, & Co. Ltd.	170
Sanders, W. H., (Electronics) Ltd.	498
Scott, Geo. L., & Co. Ltd.	50
Semiconductors Ltd.	310
Service Electric Co. Ltd.	356
Sheffield Steel Products Ltd.	483
Simmonds Aerocessories Ltd.	278
Sims, F. D., Ltd.	484
Smith & Nephew Ltd.	355
South London Electrical Equipment Co. Ltd.	487
Spear Engineering Co. Ltd.	215
Stability Capacitors Ltd.	266
Standard Insulator Co. Ltd.	54
Standard Telephones & Cables Ltd.	167
Steatite & Porcelain Products Ltd.	307
Stevenage Relays Ltd.	498
Stocko (Metal Works) Ltd.	101
Stratton & Co. Ltd.	201
Suflex Ltd.	302
Swift, Levick & Sons Ltd.	258
Sylvania Thorn Colour Television Laboratories Ltd.	274
Symons, H. D., & Co. Ltd.	277
Taylor Electrical Instruments Ltd.	211
Technograph Electronic Products Ltd.	463
Tektronix Inc.	451
Telcon Metals Ltd.	306
Telegraph Condenser Co. Ltd.	205
Telephone Manufacturing Co. Ltd.	304
Telequipment Ltd.	497
Texas Instruments Ltd.	303
Thermionic Products (Electronics) Ltd.	106
Thermo-Plastics Ltd.	209
Thorn Electrical Industries Ltd.	480
Tickopres Ltd.	454
Transradio Ltd.	452
Tucker, George, Eyelet Co. Ltd.	261
Turner, Ernest, (Spa Lane Mills) Ltd.	110
United Trade Press Ltd.	467
Vactite Wire Co. Ltd.	365
Venner Electronics Ltd.	496
Vidor Ltd.	52

(Continued on page 210)

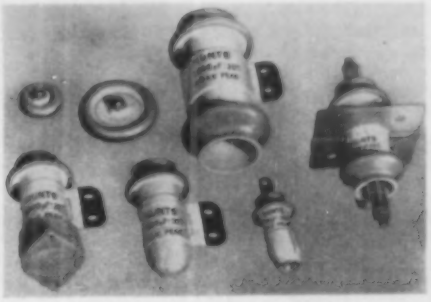
With the increasing production of analogue computers, a demand for high-resolution potentiometers of small physical size has been created. In many cases, helical potentiometers satisfy the demands and, therefore, a number of manufacturers are developing smaller and smaller units. One such unit is the helical potentiometer type HEL.07-10 which is produced by Reliance Manufacturing. This is a 10-turn potentiometer measuring 2 in. long × ¾ in. diam., and standard models are produced with resistance values ranging from 100 Ω to 30 kΩ with a tolerance of ±5%. Accuracy figures are within ±1% on the standard models; specials can be supplied to within 0·25%. Reliance have also developed multi-gang helical potentiometers in synchro size 11



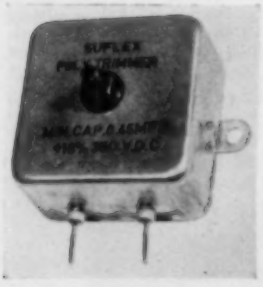
Among the many items being shown on the S.T.C. stand is a range of small l.f. quartz crystals for use in miniature equipments. These units are in evacuated glass envelopes of 0·4 in. diameter and have wire leads. The length of the units is dependent on frequency and varies from 2·83 in. for the frequency range 57–62·9 kc/s to 1·45 in. for the range 100–150 kc/s. Accuracy is ±0·01% at 24 °C and stability is ±0·2% from 70 °C to −20 °C. It is intended to extend the frequency range of this type down to 10 kc/s in future to replace types using the current B7G mounting



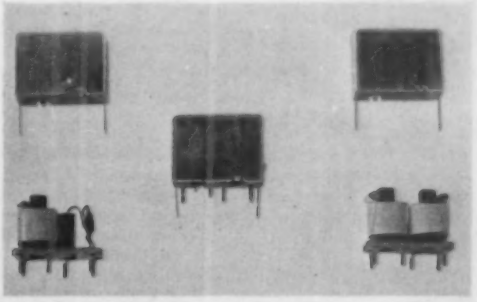
The illustration shows one of the many plug-and-socket arrangements which have been developed for printed-circuit board mounting. This is a Belling & Lee sub-miniature coaxial connector (the picture left shows an exploded view of the plug, while the picture right shows the complete plug and socket)



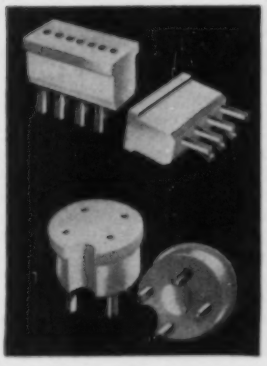
One of the latest developments to be shown on Hunt's stand is a range of ceramic power capacitors. This range is produced in tubular, disc and pot form and the tubular-construction types are available from 750–2,000 pF for 8·5-kV peak working (max. current 50 A); the pot-construction types cover the capacitance range 200–1,200 pF, and the disc types cover the capacitance range 15–500 pF



A trimmer of unusually large capacitance value is being shown by Suflex. This is a polystyrene capacitor with a continuously-variable range from 0·45 μF to 0·5 μF and is mounted in a hermetically-sealed case measuring 2 in. × 2 in. × 1¼ in. The adjustment is by means of an Allen screw and the accuracy is within 0·1%. The complete range of these trimmers includes units with a 10% variation of capacity



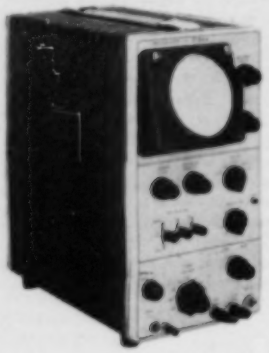
The increasing demand for personal portable radios is reflected in the development of sub-miniature components of all kinds including i.f. transformers and coils. Examples of the present-day trend in sub-miniature coils are shown by Brayhead. One of these is a bandpass i.f. transformer which uses side-by-side coil mounting, with the advantage that adjustment of tuning cores can be carried out from the top of the can. The can, which measures 22 mm × 12½ mm × 19 mm, is also used to house a final i.f. transformer complete with a diode and filter capacitor



Holders which facilitate the ready replacement of transistors are being increasingly used in specialized equipment and therefore a number of different types are being produced. Sealectro are making a p.t.f.e. type which uses a 'press-fit' method of fixing which requires holes to be drilled or punched in the chassis and the holder is pressed into position with a simple tool. The contacts are gold-plated beryllium copper and the 'pin' sockets are formed from rectangular channel section which is suitable for transistor leads of from 0·01 in. to 0·024 in. diameter

R.E.C.M.F. Show 1961 Continued

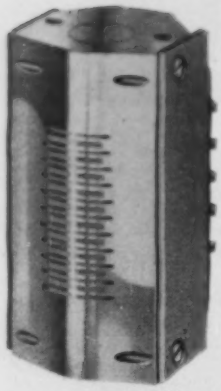
EXHIBITORS	STAND NO.
Wandleside Cable Works Ltd.	**229**
Wayne Kerr Laboratories Ltd.	**485**
Wego Condenser Co. Ltd.	**203**
Welwyn Electric Ltd.	**260**
Westinghouse Brake & Signal Co. Ltd.	**153**
Weymouth Radio Mfg. Co. Ltd.	**218**
Whiteley Electrical Radio Co. Ltd.	**56**
Widney Manufacturing Co. Ltd., The	**224**
Wingrove & Rogers Ltd.	**217**
Wire Products & Machine Design Ltd.	**490**
Wireless & Electrical Trader	**489**
Wireless Telephone Co. Ltd.	**310**
Wireless World	**489**
Woden Transformer Co. Ltd.	**267**
Wolsey Electronics Ltd.	**405**
Zenith Electric Co. Ltd.	**257**



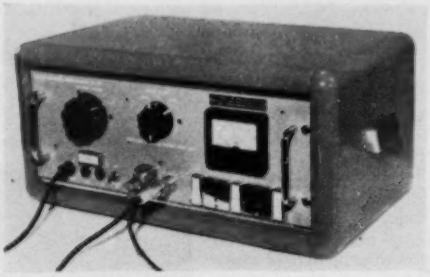
Telequipment have introduced the S42A Serviscope, a wide-range development of the S42, with an increase in the Y-amplifier bandwidth to 9 Mc/s at the normal gain of 100 mV/cm while preserving all the features of the earlier model, including the × 10 gain facility at limited bandwidth (illustrated left)

A fully-encapsulated transistorized chopper with no moving parts is being displayed by T.M.C. This is model SM10 and it is designed for low-level d.c. chopping. It has a built-in wave-shaping circuit, will operate from any 6-V sinusoidal or square-wave supply from 40 c/s to 1 kc/s, and will switch a range of chopping voltages from 1 μV to 250 mV

Another new instrument introduced at the Show is the Marconi Precision Crystal Calibrator Type TF1374. This covers the frequency range 100 kc/s to 1,500 Mc/s with a stability of 1 part in 10^6 per month, and provides precision crystal check points in the form of headphone zero-beats. The points occur at switch-selected intervals of 10 Mc/s, 2 Mc/s, 500 kc/s or 10 kc/s. Each series of points is available separately and completely free from sub-harmonic beats (illustrated above). The 100-kc/s points are clear and distinct up to at least 500 Mc/s and the 10-kc/s points are usable up to 10 Mc/s

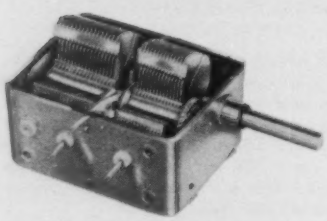


To provide a multi-channel recording equipment which could be assembled to suit particular applications, Thermionic Products are making a number of sub-units, such as a multi-channel tape deck, record-amplifier unit, monitor and replay unit. Illustrated is their 10-channel record head stack which is designed for a.f. recording 20 tracks on a 1-in. magnetic tape by interleaving two 10-head stacks



Designed primarily to provide a rapid means for carrying out production voltage-proof tests on components, the LEMCO Impulse Voltage-Proof Test Set is one of the many instruments which have been specially produced for the components industry. With this instrument, test conditions are such that a d.c. voltage-proof test of 1-min. duration is simulated by a train of impulses of a few milliseconds duration. The output is continuously variable up to a maximum equivalent of 2 kV d.c. Panel lights are provided for pass-and-fail indication, along with counters which record the pass-and-fail totals. There are connections for additional indicator lamps and for driving an external relay when the instrument is used for automatic testing

Indicative of the modern trend in the construction of variable capacitors is the Plessey type 28. This is robustly and accurately made to eliminate microphony and to achieve a high order of capacitance matching between the two sections. In the standard model, a step-down gear arrangement is fitted with a ratio of 3 : 1; the gears are spring-loaded to eliminate back-lash and are also available with ratios of 6 and 8 : 1. The supporting frame dimensions are approximately 2½ in. × 2 in. × 1½ in. and the maximum capacitance is 486 pF for the r.f. section and 430 pF for the oscillator section. Power factor at 1 Mc/s at maximum capacitance is less than 0·001



Lens-Compensated Biconical Aerial

DESIGN CRITERIA

By L. Solymar*

SUMMARY. *The dimensions of a biconical aerial can be significantly reduced by correcting the phase error in the mouth by a hyperbolic lens. The radiation pattern of this aerial is calculated with the aid of the Stratton-Chu formula.*

A practical example is worked out to show the design procedure if the half-power points are given.

Properties of the biconical aerial were investigated a long time ago[1]. Its simple geometrical shape made possible the solution of the boundary value problem, which played an important role in clarifying the concepts of aerial theory. However, in spite of its advantages (wide frequency band and considerably more directive than the dipole) its practical application has been rather limited. At longer wavelengths its weight is excessive while, at shorter wavelengths, slot aerial arrays proved to be more useful.

Nevertheless, if in the mouth of the biconical horn the phase error is corrected by a lens, its performance is comparable with that of other omnidirectional aerial arrays.

In the present article, the radiation pattern of this lens-compensated biconical aerial is calculated. Our method of solution is to calculate first the electric and magnetic fields in the mouth of the biconical horn and then to apply the integral formulae derived by Stratton and Chu[2].

The Electric Field in the Mouth of the Biconical Horn

The cross-section of the aerial may be seen in Fig. 1. Since it is fed by a coaxial waveguide, a pure TEM mode is assumed to propagate in the horn. The outer surface of the dielectric lens used is cylindrical while the inner surface is a hyperboloid. The role of the lens is to convert the spherical wave into a cylindrical one.

However, as the phase front is corrected by the lens, the amplitude distribution is distorted at the same time. The resultant distribution along the z axis is given by the following formula[3].

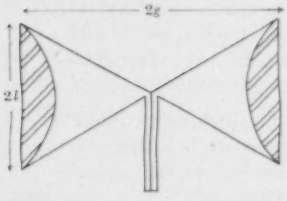


Fig. 1. *A cross section of the biconical aerial*

$$A(z) = \frac{1}{f^4} A(\alpha) \frac{n \cos \alpha - 1}{\sqrt{(n-1)(n-\cos\alpha)}} \quad .. \quad (1)$$

where

α = the angle between the vector $\boldsymbol{u}$ and the xy plane (Fig. 2).

$\boldsymbol{u}$ = a vector pointing from the centre of the aerial (denoted by 0 on Fig. 2) to an arbitrary point on the aperture.

$A(\alpha)$ = the angular distribution of the incident wave, which in the present case is proportional to $\cos \alpha$.

n = the refraction index of the lens.

f = the focal length of the lens.

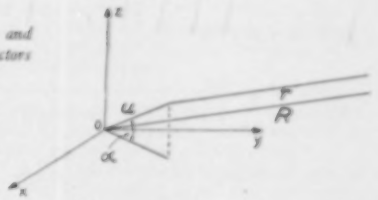


Fig. 2. *Shows angle α and vector $\boldsymbol{u}$ [Equ. (1)] and vectors r and R [Equ. (6)]*

If ρ is large in comparison with the wavelength the magnetic field in the aperture can be expressed by the simple plane-wave relationship.

Summarizing, the electric and magnetic fields in the aperture are as follows:

$$\boldsymbol{E} = E_0 A(z) \boldsymbol{i}_x, \quad \boldsymbol{H} = (\boldsymbol{n} \times \boldsymbol{E}) \sqrt{\frac{\epsilon}{\mu}}$$

where $\boldsymbol{n}$ is the unit vector normal to the aperture directed outward.

Calculation of the Electric Intensity in the Fraunhofer Region

The electric field at a point P can be expressed from the fields existing on the bounding surfaces. In the present case, the bounding surfaces are infinity (where the field is zero) and the aperture of the biconical horn.

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Thus the electric field at the point P is given by the following formula[4]

$$\boldsymbol{E}_p = \frac{1}{4\pi}\int[-j\omega\mu\,\psi\,(\boldsymbol{n}\times\boldsymbol{H}) + (\boldsymbol{n}\times\boldsymbol{E})\times\nabla\psi + (\boldsymbol{n}\,\boldsymbol{E})\,\nabla\psi]d\rho \quad .. \quad .. \quad .. \quad (4)$$

where

ρ = the aperture of the biconical horn.

ω = the angular frequency.

μ = the magnetic permeability.

$\psi = \frac{e^{-j\beta r}}{r}$

r = the distance from P to any point on the aperture.

Performing the operations and neglecting terms higher than $1/r$, we obtain the electric intensity at the distant point P as follows,

$$\boldsymbol{E}_P = \frac{j\beta}{4\pi}\int\frac{e^{-j\beta r}}{r}\{[1+(\boldsymbol{i}_r\,\boldsymbol{n})]\,\boldsymbol{E} - (\boldsymbol{i}_r\,\boldsymbol{E})\,\boldsymbol{n}\}d\rho \quad (5)$$

where $\boldsymbol{i}_r$ is a unit vector pointing from the aperture in the direction of the point P.

Since the point P is very far, the $\boldsymbol{OP} = \boldsymbol{R}$ can be taken parallel to $\boldsymbol{r}$ (Fig. 2). Thus,

$$r = R - \boldsymbol{u}\,\boldsymbol{i}_r \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

Expressing the vectors in Equs (5) and (6) with the elevation angle θ and azimuth angle ϕ and introducing furthermore the variables

$$v = \beta\,l\sin\theta,\ s = \beta\rho\cos\theta,\ \zeta = \frac{z}{l},\ A(z) = F(\zeta) \quad (7)$$

we get

$$\boldsymbol{E}_p = j\,\frac{\beta\rho l}{2}\,\frac{e^{-j\beta R}}{R}\,E_0\,g(v)\,\boldsymbol{N}(s) \quad .. \quad .. \quad .. \quad (8)$$

where

$$g(v) = \int_{-1}^{1} F(\zeta)\,e^{jv\zeta}\,d\zeta \quad .. \quad .. \quad .. \quad .. \quad (9)$$

$$\boldsymbol{N}(s) = \frac{1}{2\pi}\int_0^{2\pi} e^{js\cos\phi}\{[1+\cos(\phi-\theta)]\,\boldsymbol{i}_z - \sin\theta\,\boldsymbol{n}\}\,d\phi$$
$$= [J_0(s) + j\cos\theta\,J_1(s)]\,\boldsymbol{i}_z - \sin\theta\,J_0(s)\,\boldsymbol{n} \quad .. \quad .. \quad (10)$$

The normalized power radiation pattern can be expressed from Equ. (8) as follows:

$$p(\theta) = \frac{g^2(v)}{g^2(0)}\,\frac{|\boldsymbol{N}(s)|^2}{|\boldsymbol{N}(\beta\rho)|^2} \quad .. \quad .. \quad .. \quad .. \quad (11)$$

As can be seen from Equ. (9) the first factor in Equ. (11) is the familiar radiation pattern of a line source, the distribution function of which is $F(\zeta)$. However, in the present case due to the cylindrical aperture a correction factor $|\boldsymbol{N}(\rho)|^2/|\boldsymbol{N}(\beta\rho)|^2$ appears. This factor depends only on ρ/λ and θ. Its value is very close to unity for small values of θ but violently oscillates for larger values as it is shown in Table 1. Since in most practical cases this affects only the distant side lobes,

TABLE 1

$\theta°$ / ρ/λ	0	2	4	6	8	10	12	14
4	1·000	1·000	1·000	1·000	1·000	0·994	0·983	0·970
6	1·000	1·000	1·000	1·000	0·997	0·987	0·978	0·983
8	1·000	1·000	1·000	0·999	0·994	0·985	0·986	1·021
10	1·000	1·000	1·000	0·998	0·991	0·986	1·005	1·064

$\theta°$ / ρ/λ	16	18	20	22	24	26	28	30
4	0·967	0·988	1·050	1·153	1·264	1·311	1·230	1·029
6	1·022	1·104	1·187	1·186	1·052	0·904	0·991	1·335
8	1·094	1·150	1·091	0·948	0·973	1·258	1·311	0·928
10	1·118	1·065	0·944	1·037	1·274	1·080	0·923	1·416

the second factor can be generally disregarded for the design of the aerial. Thus the main lobe is determined by the distribution of the electric intensity along the z direction.

Evaluation of the $g(v)$ Function

If the dielectric material of the lens is given then with the aid of Equs (1), (7) and (9) $g(v)$ can be calculated. In most practical cases there is no need to apply numerical methods since $F(\zeta)$ can be well approximated by the following polynomial

$$\sqrt{f}F(\rho) = \sum_{k=0}^{m} a_{2k}\left(\frac{l}{\rho}\right)^{2k}\zeta^{2k} \quad .. \quad .. \quad .. \quad (12)$$

where the value of m is generally 2 or 3.

Substituting Equs. (12) into Equ. (9) and performing the integration we get

$$\frac{g(v)}{g(0)} = \frac{\frac{1}{v}\sum_{k=0}^{m}(-1)^k\left(\frac{l}{\rho}\right)^{2k}\frac{a_{2k}\,(2k)\,!}{v^{2k}}\,[C_{2k}(v)\sin v - S_{2k}(v)\cos v]}{\sum_{k=0}^{m}\frac{a_{2k}}{2k+1}\left(\frac{l}{\rho}\right)^{2k}} \quad .. \quad .. \quad (13)$$

where

$$C_{2k}(v) = \sum_{i=0}^{k}\frac{v^{2i}}{(2i)\,!}(-1)^i \quad .. \quad .. \quad .. \quad (14)$$

and

$$S_{2k}(v) = \sum_{i=1}^{k}\frac{v^{2i-1}}{(2i-1)\,!}(-1)^{i-1} \quad .. \quad .. \quad (15)$$

Example

For a lens of polystyrene or plexiglass $n = 1{\cdot}6$ and the coefficients of the polynomial approximating $\sqrt{f}F(\zeta)$ are

$$a_0 = 1,\ a_2 = -2{\cdot}184,\ a_4 = 2{\cdot}040$$

In Figs. (3) and (4) $[g(v)/g(0)]^2$ is plotted against

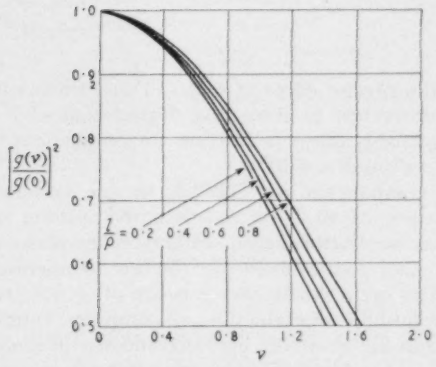

Fig. 3. $[g(v)/g(0)]^2$ *plotted against* v *for various values of* l/ρ

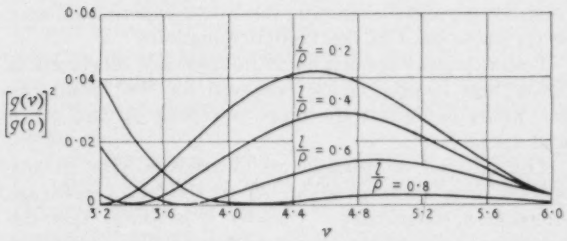

Fig. 4. *While Fig.* 3 *shows the main lobe down to half-power points this figure shows the region of the first side-lobe*

v for $l/\rho = {\cdot}2, {\cdot}4, {\cdot}6, {\cdot}8$. Fig. (3) shows the main lobe down to the half-power points, while the region of the first side lobe is shown (in a different scale) in Fig. (4). It can be seen that the larger the value of l/ρ the wider is the width of the main lobe and the smaller is the level of the side lobes. This is simply the consequence of the more tapered distribution function.

The value of l can be determined from the prescribed half-power angle while the choice of ρ is influenced by the following considerations.

(1) Large value of l/ρ reduces the size of the aerial but at the same time it increases the beam-width and makes the lens thicker which, on the other hand, increases the losses and the weight.

(2) Small value of l/ρ makes possible the use of a thin lens and increases the aperture efficiency but makes the size excessive.

Obviously a compromise is needed which can be achieved by choosing $l/\rho \simeq 0{\cdot}5$.

If the half-power angle is 5 degrees, then

$$l/\lambda = \frac{v_{\frac{1}{2}}}{2\pi \sin \theta_{\frac{1}{2}}} = 2{\cdot}78,\ \rho/\lambda = 5{\cdot}56$$

The thickness of the lens can be calculated from the equation of the contour[3].

$$\frac{d}{\lambda} = \frac{1}{n-1}\left[\sqrt{1 + (l/\rho)^2} - 1\right]\frac{\rho}{\lambda} = 1{\cdot}1,$$

with which all the dimensions of the aerial are determined.

REFERENCES

[1] S. A. Schelkunoff, "Electromagnetic Waves", D. Von Nostrand, 1943.
[2] J. A. Stratton and L. J. Chu, *Phys. Rev.*, 1939, Vol. 56, p. 99.
[3] S. Silver, "Microwave Antenna Theory and Design", McGraw-Hill, pp. 390–391, 1949.
[4] ibid, p. 83.

SEMICONDUCTOR WAFERS POLISHED ELECTROCHEMICALLY

A new electrochemical technique for rapid scratch-free polishing of germanium and silicon wafers for transistors has recently been announced by Bell Telephone Laboratories. The new polishing method is claimed to be much faster and more efficient than conventional polishing methods.

In conventional manufacturing practice, semiconductor material slices are prepared in four steps. First, they are cut from a single-crystal rod, then they are lapped with a coarse abrasive and polished on optical lapping machines. These three operations produce a smooth flat surface. In the final step, the slices are etched to remove all the residual mechanical damage to the crystal face caused by abrasive action.

When there are deep scratches from lapping and polishing, the etchant cannot remove all the damage. In contrast, electrochemical polishing does not introduce new damage during processing.

In the technique announced, semiconductor material slices are mounted on a non-conducting disc. After electrical contact is made to the slices, they are placed on a polishing wheel over which an electrolyte flows. When the polishing wheel rotates, a film of electrolyte whose thickness is determined principally by its viscosity separates and automatically maintains the semiconductor at a relatively constant distance from the wheel.

Dilute potassium hydroxide is the electrolyte used for the electro-polishing of germanium and dilute hydrofluoric acid is used for silicon. It is reported that for n-type materials, it is desirable to illuminate the semiconductor surface during electroetching.

Semiconductor slices polished by this new technique show no texture that can be associated with surface roughness even under electron-microscope examination at a magnification of 53,000 times, whereas a mechanically-polished surface shows distinct scratches when magnified 500 times.

The illustration shows the equipment used for electrochemical polishing. In operation, the electrolyte (in the glass container) drops on to the rotating polishing wheel

Electronic Components for Computers

DETERMINATION OF FAILURE RATES

By L. Knight, A.M.I.E.E.*

A computer may contain 100,000 or more electronic components and, if it is to be reliable, the average failure rate of all these components must be small. Relevant information on the reliability of some components exists in the form of field experience with other equipment, but it is necessary to implement this with laboratory tests to obtain assurance that the reliability of the newer components will be adequate.

In a computer, a component is said to have failed when one or more of its parameters have changed to an extent which causes malfunctioning of the circuit concerned. Such a change may be the result of gradual deterioration with age, in which case it would be referred to as a 'degradation' failure. Alternatively, there may be a quite sudden and drastic change in the electrical characteristics, possibly to the extent of an open-circuit or short-circuit condition, in which case it would be termed a 'catastrophic' failure. For the purpose of reliability studies, it is often useful to consider these two types of failure separately.

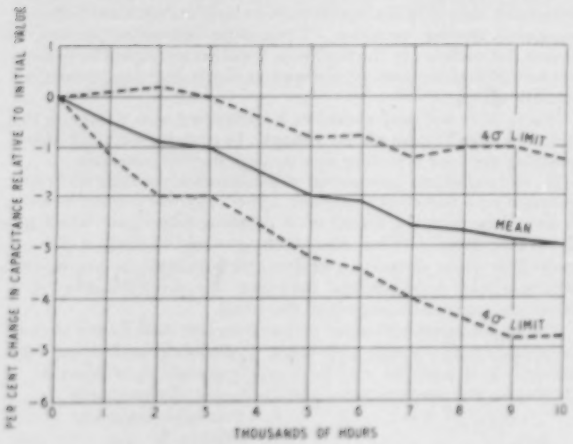

Fig. 1. Capacitance change of 'Brand-X' capacitors operated at 40 °C with 50 volts d.c. applied and no a.c. ripple

Degradation Failures

The degradation experienced by any component is largely dependent on the way in which it is used. The nature of the imposed electrical stresses, the temperature cycling and the humidity cycling all play a part and it would be a rather formidable task to isolate the precise effect of each. Thus, when running a laboratory test to study the degradation of a component, care is taken to operate the component under typical computer conditions.

The components are placed in an ambient air temperature of 40 °C to simulate the interior of the computer. Furthermore, rather than maintain a precise and constant 40 °C, the air temperature is allowed to cycle slowly over a range of ± 5 °C to give a more faithful reproduction of computer conditions. Computers are normally operated indoors in temperate climates. It is therefore assumed that the humidity conditions obtaining in the laboratory will be similar to those experienced by an actual computer. Some computers, in fact, are sent to tropical climates but, in such cases, the computer rooms usually have some degree of air-conditioning and the climatic conditions experienced may be no worse than in the United Kingdom.

Tests normally operate for 20 hours a day, all electrical power and heating being removed for the remaining four hours to imitate the daily switching on and off of a computer.

The number of components of any one type in any one test may be from about 30 to 100 and results are interpreted statistically. Usually it is found that the parameter with which one is concerned has a Gaussian distribution, or something closely approaching one. In such a case the results would be plotted as in Fig. 1, which gives the drift in capacitance of a hypothetical 'Brand-X' capacitor.

The solid line represents the mean value of capacitance change and the dotted lines represent ($\pm 4\sigma$) about the mean. In theory, one can expect only about one capacitor in 15,000 to fall outside the 4σ limits. Thus one can deduce that when using Brand-X capacitors under the conditions specified, and provided that the circuits will accept capacitance drifts not exceeding $+\frac{1}{2}\%$, -5%, there is not likely to be more than one degradation failure per 15,000 capacitors in 10,000 hours.

Sometimes it is useful to consider the effects of temperature coefficients and voltage coefficients jointly with the effects of ageing. Fig. 2 gives the results of a typical resistor test treated in this way.

The upper solid line indicates the mean resistance change (from the original value at 25 °C and with no dissipation) as measured when the resistors were dissipating 0·4 watt and in an ambient temperature of 50 °C. The lower line represents the mean resistance change with the resistors measured in an ambient temperature of 0 °C and with no dissipation.

* *Reliability Department, I.C.T. (Engineering) Ltd.*

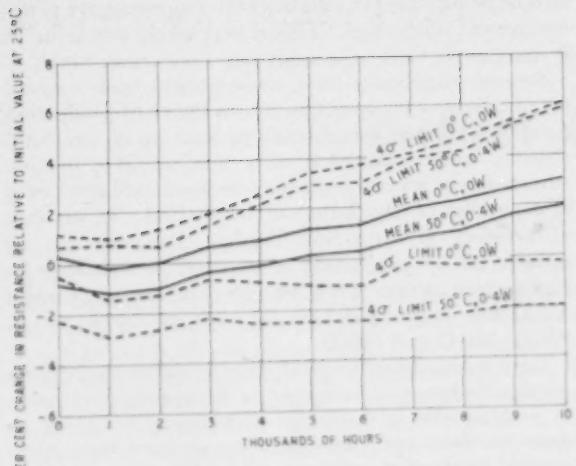
PER CENT CHANGE IN RESISTANCE RELATIVE TO INITIAL VALUE AT 25°C
4σ LIMIT 0°C, 0W
4σ LIMIT 50°C, 0·4W
MEAN 0°C, 0W
MEAN 50°C, 0·4W
4σ LIMIT 0°C, 0W
4σ LIMIT 50°C, 0·4W
THOUSANDS OF HOURS

Fig. 2. *Resistance change of type XYZ resistors normally operating at 40 °C, 0·4-W dissipation*

The dotted lines represent the 4σ limits outside the means. Thus one can deduce that if the ambient temperature is always between 0 °C and 50 °C and, if the dissipation is between 0 and 0·4 watt, the deviation in resistance is unlikely to exceed the limits described by the outer dotted lines.

It is not essential to take the 4σ limits. These were chosen because it was considered desirable to limit the number of degradation failures to less than 0·01% of the components. If a higher percentage of degradation failures were permissible, 3σ or 3½σ limits might be used.

With most components it is practicable to design the circuits to accept the worst degradation which is likely to occur during the lifetime of the computer but, with thermionic valves, added difficulties arise. To illustrate this, Fig. 3 gives the results of a life test on one type of valve. The curve plots anode current under certain circuit conditions and it will be seen that if the number of degradation failures is to be kept low over, say, 15,000 hours the circuits must tolerate a very serious deterioration in anode current. In practice, a designer would compromise and design for say 50% deterioration. The curves in Fig. 3 would then be used to indicate the optimum time at which to replace the valves.

In some cases, it is some function of a component parameter which is found to have a Gaussian distribution, not the parameter itself. For example, with transistors it is found that it is the logarithm of collector–base leakage current, not the collector–base leakage current itself, which has a Gaussian distribution. Thus when analysing results it would be the logarithm of the current which would be plotted.

Catastrophic Failures

Catastrophic failures are often due to odd mechanical defects in manufacture but may sometimes occur as a result of the normal slow degradation having been accelerated by some special circumstances. In general, one cannot make any statistical extrapolation from a small sample and, if one wants to verify that a component has a catastrophic failure rate less than 0·01% per 1,000 hours, there is usually no alternative but to embark on a test involving at least 25,000,000 component-hours. Whether this consists of 2,500 components for 10,000 hours or 10,000 components for 2,500 hours it is quite an expensive test.

There is a temptation to try and accelerate the test. For example, one might argue that since the speed of chemical reactions tends to double for each 10 °C rise in temperature, catastrophic failure rates are likely to follow the same law. One might therefore assume that a failure rate of 0·16% per 1,000 hours at 80 °C was equivalent to a failure rate of 0·01% per 1,000 hours at 40 °C. However, the use of such deductions to extrapolate information from accelerated tests is felt to be extremely risky.

In general, it is safer to perform a modest test under simulated computer conditions which will give one reasonable confidence that the catastrophic failure rate is below, say, 0·1% per 1,000 hours rather than embark on an accelerated test which may give a very doubtful 'proof' that the failure rate is better than 0·01%.

Very often, even a modest test will show up some manufacturing defects which are likely to impair the reliability of a component. Such defects might include a tendency for end caps to be loose, a tendency for internal welds to become open-circuit, poor solderability of the wire ends, etc. Even if the test does little more than to highlight these and initiate corrective action at the manufacturers' plant the time and expense is well repaid.

One case in which the results of accelerating a test seem to be reasonably predictable is that of capacitors. Evidence of several types indicates that the failure rate is related to the applied voltage in the way shown in Fig. 4. Thus, if the catastrophic failure rate at 1·5 times the rated working voltage is found to be 10% per 1,000 hours, one can assume that the failure rate at 0·2 of the rated voltage will be of the order of 0·015% per 1,000 hours.

Another case in which the results of acceleration are reasonably predictable is that of connectors. The main effect of ageing is the corrosion of the mating surfaces

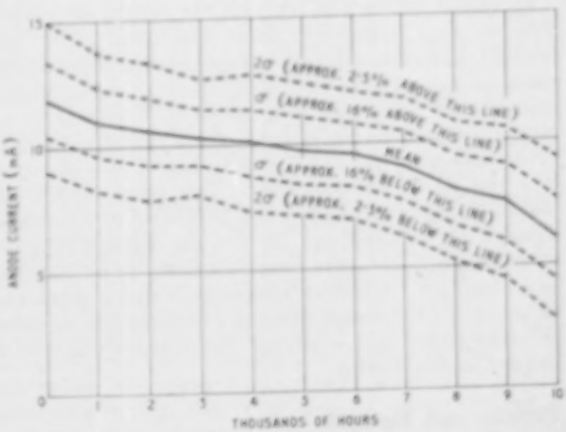
ANODE CURRENT (mA)
2σ (APPROX. 2·5% ABOVE THIS LINE)
σ (APPROX. 16% ABOVE THIS LINE)
MEAN
σ (APPROX. 16% BELOW THIS LINE)
2σ (APPROX. 2·5% BELOW THIS LINE)
THOUSANDS OF HOURS

Fig. 3. *Anode current of a particular valve type when operated in given circuit conditions*

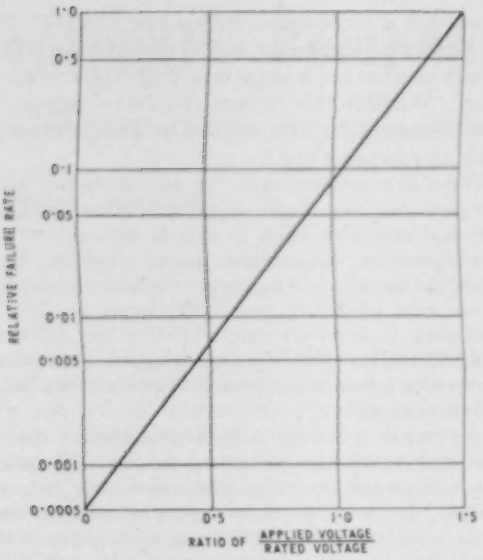

Fig. 4. *Relationship between the catastrophic failure rate of a capacitor and the applied voltage. This relationship appears to be approximately correct for a number of paper, polystyrene and ceramic types*

due to atmospheric contamination. Thus connectors are tested in a chamber in which is burnt coal gas to which a trace of carbon disulphide has been added. This pollutes the atmosphere inside the chamber with sulphurous gases to an extent some 20 times worse than that found in a bad industrial atmosphere and is therefore assumed to cause corrosion 20 times as fast.

Test Boxes

Most of the boxes used for testing components are, for convenience, built around standard relay racks. Air is continually circulated up the back of the rack and down the front, a heater and thermostat controlling the temperature. The temperature cycling about the mean is achieved automatically by the backlash in the thermostat used.

In earlier versions of test box the components were held in clips improvised from hair clips. For the periodic measurements each component was individually removed and taken to the test equipment. An improved method now being used is to mount the components on plug-in printed-wiring boards. Each board holds 6 to 12 components and the complete board is transferred to a test jig for measurement. This latter system has the advantage that there is less likelihood of accidental damage due to handling. In fact, the damage incurred by handling is no worse than that likely to be incurred in a computer. A further advantage is that a check can be made of any possible damage which may be incurred by assembling and soldering the components on to printed wiring boards.

Pitfalls and Hazards

When performing a test to study degradation or catastrophic failures it is extremely important that the sample being tested should be really representative of the component concerned. This is very likely not to be so if the sample only represents one production batch.

Several techniques have consequently been evolved for obtaining a cross-section of a number of production batches. The test sample may be built up of, say, half-a-dozen units from each of a number of deliveries. Alternatively, several units may be purchased from each of a number of distributors or wholesalers. In the case of resistors or capacitors further diversion can be introduced by purchasing a selection of resistance or capacitance values. Thus, instead of 50 resistors each of 470 Ω one might purchase 10 each of 390 Ω, 430 Ω, 470 Ω, 510 Ω and 560 Ω.

Care is also taken to avoid placing too much reliance on the evidence of prototype or first-production units. If nothing else is available, preliminary tests will be done on these and confirmatory tests done later when normal production units are available.

Experience has shown that it is essential to judge the performance of a component in relationship to the operating conditions and the circuits concerned. Unfortunately the situation is never so simple that one can say that 'Brand-X' capacitors will always have a failure rate of 0·02% per 1,000 hours no matter how they are used. Nor is it even so simple that one can say that 'Brand-X' capacitors will always give a lower failure rate than 'Brand-Y' capacitors.

As a particular example which illustrates this point it is interesting to consider the results obtained with commercial-grade 12AU7 valves. Tests on Make 'A' showed very little deterioration in performance when the valves were operated with a fairly high anode current but a severe deterioration when they were operated with a very low anode current. Make 'B' on the other hand, showed some deterioration when operated at a high anode current but very little extra deterioration when operated with a very low anode current. It so happened that, because of the peculiarities of the circuits concerned, Make 'A' was far superior for one range of computing equipment but Make 'B' was far superior for another range of equipment!

Conclusions

Even extensive laboratory tests cannot give complete confidence that the performance of electronic components will be adequate when actually in use in a computer. Nevertheless, it has been found that modest tests such as outlined do give a welcome degree of assurance and allow many potential defects in components and component usage to be corrected during the design stage. Using old-fashioned wait-and-see techniques, many of these defects might not have been found until several production computers were in the field.

INTERNATIONAL PLASTICS EXHIBITION

The sixth International Plastics Exhibition and Convention is being held at Olympia, London, from 21st June to 1st July. It is organized for *British Plastics* and *International Plastics Engineering* by Iliffe Exhibitions Ltd. with the co-operation of the British Plastics Federation.

The Exhibition is to be opened this year by Lord Hailsham, the Minister for Science.

Critical Comparison

In the May 1960 'Tool', we noted that when it is necessary to decide whether or not a shaft of diameter d_1 will enter a bush of diameter d_2, the important question is the sign of the quantity $(d_2 - d_1)$, not its size. If $d_1 \geqslant d_2$, the shaft is useless for the purpose, however small the difference $d_1 - d_2$ may be. If $d_1 < d_2$ as measured by a suitable gauge, then the shaft will enter the bush and is to be regarded as completely satisfactory from this point of view however small $(d_2 - d_1)$ may be. There may be some other restriction, that d_1 must exceed another quantity d_0, but if so, this is a completely separate issue and there is no point in applying a second gauge to test whether d_1 exceeds d_0 or not except to those shafts for which d_1 is already known to be less than d_2. As long as d_1 is between d_0 and d_2, the shaft will do; we are not interested in the values of $(d_1 - d_0)$ or $(d_2 - d_1)$ in cases of this kind.

For purposes of calculation, it is admittedly easier to handle numbers as if they were exact and, in the June 1960 'Tool', we saw that this was quite legitimate as long as we are always on the lookout for the cardinal sin of mistaking a metrical zero for a logical one, and then making matters worse by allowing this erroneous logical zero to get into the denominator (or a logarithm or similar expression arising from the integration or manipulation of such a denominator). It is essential to remember that metrical numbers are never exactly on a boundary.

Sometimes it is not particularly important what the value or sign of a tolerance is; this happens when no critical comparison is made, and is unlike the shaft and bush case. Suppose, for example, we have a 6-V battery with a 10-Ω resistor connected across it, and that the tolerance associated with both is ± 10%. Then, if we ignore the tolerances altogether, we find that the current is 0·6 A. Extreme values of current would occur if the maximum voltage 6·6 V occurred together with the minimum resistance 9 Ω, or if the minimum voltage 5·4 V occurred together with the maximum resistance 11 Ω; thus the maximum possible current is 6·6/9 = 0·733 A and the minimum possible current is 5·4/11 = 0·491 A, and the value 0·6 A, obtained when tolerances are neglected, is approximately (but not exactly) midway between these extremes. The tolerance associated with the 0·6 A can be taken as from 18·2% below 0·6 A to 22·2% above 0·6 A. Thus, in this case, we obtain a reasonably central value of the current, 0·6 A, by neglecting tolerances, and we find that the percentage tolerance associated with the current is about double that associated with the battery voltage and the resistor which gave rise to that current. A reduction of the tolerances associated with the battery voltage and with the resistor would correspondingly reduce the tolerance associated with the current, but 0·6 A would remain a reasonably central value of that current. In this case, therefore, if we only want to know the sort of size of current to be expected, we can find this without worrying about tolerances, but the tolerances must be considered to determine the reliability or probable variation of that current.

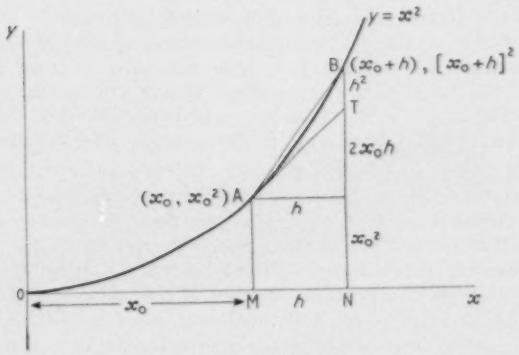

Fig. 1. *Chord and tangent of* $y = x^2$

In engineering, however, a more typical case occurs when two parts have to be fitted together, and the suitability of a particular part is tested by means of 'go' and 'not go' gauges. Here the dimensional difference between a part which will 'go' and one which will not may be extremely small but is critically important. Indeterminate cases do not arise. A difference which matters can be observed; a difference which cannot be observed does not matter and is not of critical importance. A statement that x is greater than a and less than b is unambiguous and verifiable even if $b - a$ may be very small, but from the engineer's point of view the statement $x = c$ is inadequate.

In mathematics, when a small dimension, say x, is critically important, it will arise in the crude form $1/x$, or $1/(y - d)$ will correspondingly occur if y is critically compared with d. Alternatively, the division may be disguised as a logarithm or similar expression associated with integration. The important point is that there is a critical difference between the appropriate function, say $f(y - d)$, for $(y - d)$ positive, however small, and for $(y - d)$ negative, however small. The ambiguous and indeterminate case arising when $y - d$ is a logical zero is avoided in the engineering workshop by the use of tolerances and gauges as already indicated, and in mathematics by the technique of limits—letting a variable approach nearer and nearer to a limiting value without ever reaching or passing that limiting value. In both engineering and mathematics, the sign of $(y - d)$

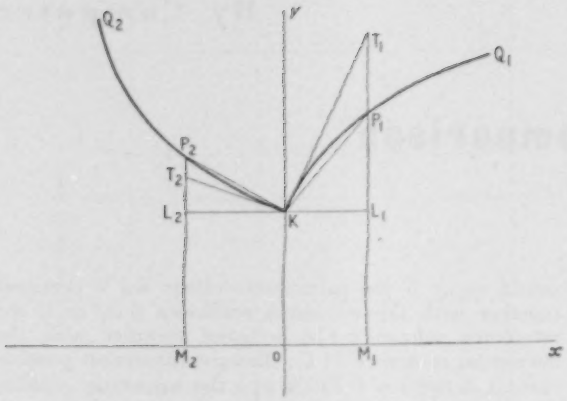

Fig. 2. Geometrical analogue of critical comparison

is clear when a limit is taken or a gauge used, though the size is not. Nearly all ambiguities in pure mathematics arise because a limit has been taken prematurely, or because the limiting process involved has not been explicitly formulated. In the engineering example above, we know that $(x - a)$ is positive and $(x - b)$ is negative, and it follows that $(x - a) < (b - a)$; the magnitude of $(x - a)$ is thus anything between 0 and $(b - a)$, but its sign is certain, and certainty of sign is the primary requirement in critical comparison.

Now suppose that the system of tolerances did not exist, and that the dimension for x was specified as a exactly. Then, if a large number of the items is made, it is reasonable to suppose that the number under size will equal the number over size, so that the probability that a particular item will be over size will be $\frac{1}{2}$. This information is useless to the engineer and still more useless to the mathematician, and yet frequently we are tempted to proceed as though some property of these items could be determined for each item, without first separating the under size items from the over size, e.g. the mean might be taken for the whole set and put forward as a reliable property of every member of the set. It cannot be too strongly emphasized that if critical comparison is involved, over size items will differ radically in some way from under size items, and should be separated as categorically as the sheep and the goats (Matthew xxv, 32). A mathematical example already considered in the August 1960 'Tool' will make this clear. If $a > 0$ and ϵ is arbitrarily small (but different from zero) and of either sign, and

$$Y_0 = \int_{-a}^{a} \frac{dx}{x} X_0 = \int_{-a}^{a} \frac{dx}{x - j\epsilon} \quad .. \quad .. \quad .. \quad (1)$$

then we there found that Y_0 was inadequately specified, while

$$X_0 = 2j \tan^{-1}(a/\epsilon) \approx j\pi \ (\epsilon > 0) \quad \text{or} \quad -j\pi \ (\epsilon < 0) \qquad (2)$$

(assuming that in Equ. (2) $\tan^{-1}\eta$ denotes the angle between $-\pi/2$ and $\pi/2$ whose tangent is η). Now, if there are 1,000 items for which X_0 has to be evaluated, and ϵ is equally likely to be positive or negative for any one item but is always small compared to a, we shall expect 500 of these items to give the value $j\pi$ for X_0 and the other 500 to give the value $-j\pi$. The mean value of X_0 for the first 500 will be $j\pi$ and the mean value for the second 500 will be $-j\pi$. If now we fail to separate the positive-ϵ items from the negative-ϵ items, we arrive at the absurd answer that the mean value of X_0 is zero—a value which never in fact occurs, since the probability that ϵ shall be a logical zero is zero, and any other value of ϵ must give either the answer $j\pi$ or the answer $-j\pi$ for each individual item, by Equ. (2). Although the sign of ϵ is thus of critical importance, the size of ϵ is not significant; as long as ϵ is sufficiently small compared to a, say less than $0{\cdot}1a$ numerically, its value only has a residual effect on the resulting value of X_0 in much the same way as the gradient $(2x_0 + h)$ of the chord AB of the curve $y = x^2$ differs by a residual amount h from the gradient $2x_0$ of the tangent AT at A in Fig. 1 when B is sufficiently close to A (h is sufficiently small). In the case of Fig. 1, however, the sign of h is not of critical importance. Fig. 2 gives a better general geometrical picture of the situation of critical comparison. Here we have the situation that when x is positive, y is related to x according to the smooth curve KP_1Q_1 (or by means of a formula from which such a smooth curve could be determined) but when x is negative, y is related to x according to a different smooth curve KP_2Q_2 (Indeed, it is not even necessary for the present discussion that the curves should meet Oy in the same point K as in Fig. 2). If x is positive, the average slope of the curve between K and a typical point P_1 near K will be a quantity P_1L_1/KL_1 which, for small x, only differs by a residual quantity from the slope T_1L_1/KL_1 of the tangent at K, whereas if x is negative, the average slope of the curve between K and a typical point P_2 near K is negative and numerically equal to P_2L_2/L_2K and this will only differ by a residual quantity from the slope (negative and numerically equal to T_2L_2/L_2K) of the tangent at K to the left-hand curve. Now the slopes of the two tangents just mentioned are radically different and their mean has no significance in relation to Fig. 2 at all. We have no right to be 'exactly' at K: it is a point which must be excluded from our discussion and consideration, although we can properly consider points like P_1 and P_2 on either side of K however close they may be to K. The probability that x is a logical zero is zero; any other value of x has an unambiguous sign and we therefore have no doubt whether it is the curve KP_1Q_1 or KP_2Q_2 which is relevant. Thus, although the point K looks somewhat odd, angular and awkward in Fig. 2, this curious look is entirely irrelevant.

In Fig. 1, the gradient of the tangent at A is the same whether A is approached from the right or from the left and, although a particularly simple curve was used in Fig. 1, this is the situation to be expected when there is no critical comparison. In the corresponding engineering case, the relevant property of each item within a suitable tolerance of a particular dimension d_0 could be determined, and the mean value taken as the measure of that property appropriate to the dimension d_0. If, however, any variation of d_0 is known or suspected to be critically important, batches of the item for which this variation is critically different must be examined separately: they are different entities having no common property.

An S-Band Circulator and Switches

The use of ferrites at microwaves is now well known and a good deal has been published about the theory of the various devices which can be constructed. The kind of performance obtainable is less familiar to most people, however, and so figures are given here to illustrate the kind of results which can be achieved. Four different devices are briefly described and the figures quoted are from measurements on experimental models developed by the Mullard Research Laboratories.

These ferrite devices are based on the phenomenon of Faraday rotation which occurs at microwave frequencies in waveguides of circular cross-section which are filled or partially filled with ferrite material, usually in the form of rods mounted coaxially with the waveguide. When a linearly-polarized wave is propagated through such a loaded waveguide in the presence of an axial magnetic field, the plane of polarization is rotated at a rate dependent upon the waveguide configuration and upon the magnetic field, but its sense is independent of the direction of the propagation and changes with the direction of the applied magnetic field.

By arranging that the angle of rotation is 45° and by terminating the ferrite-loaded section with a pair of dual-mode transducers mounted at 45° with respect to one another, a four-port element is obtained having the property that energy can be propagated non-reciprocally between pairs of ports in sequence. Such an element is known as a circulator. Since a reversal of the direction of the magnetic field results in a reversal of this sequence, a change-over switch can be based on such a circulator.

A switch can also be based on a 90° Faraday rotation. In this element, the ferrite-loaded section is terminated by a pair of junctions in line, consisting of a mode transducer for one plane of polarization and a reflectionless load for the perpendicular plane. This results in a reciprocal switch.

The crystal switch is based on the principle that when sufficiently back-biased the crystal diode behaves as a reactance whereas when forward-biased it becomes substantially resistive. Thus, by matching in the forward-bias condition, the crystal presents a reflectionless termination and by switching to the back-biased condition, the crystal presents an almost complete

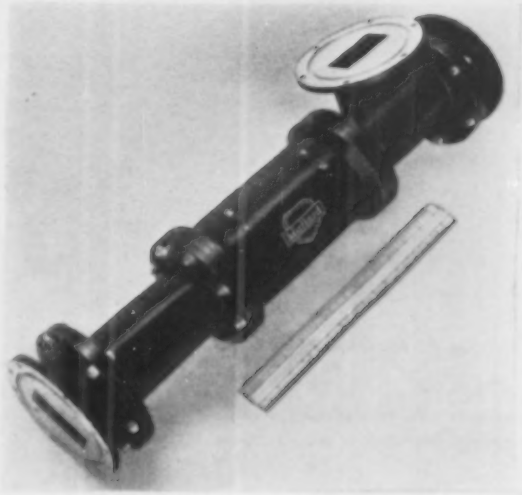

Fig. 2. *Ferrite switch, type 1, with a foot-rule alongside to indicate size*

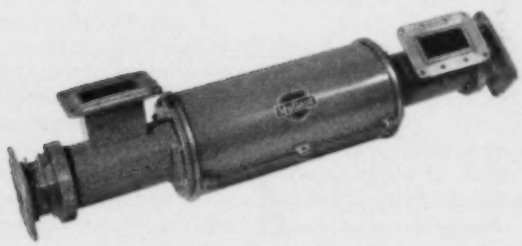

Fig. 1. *General view of complete static circulator*

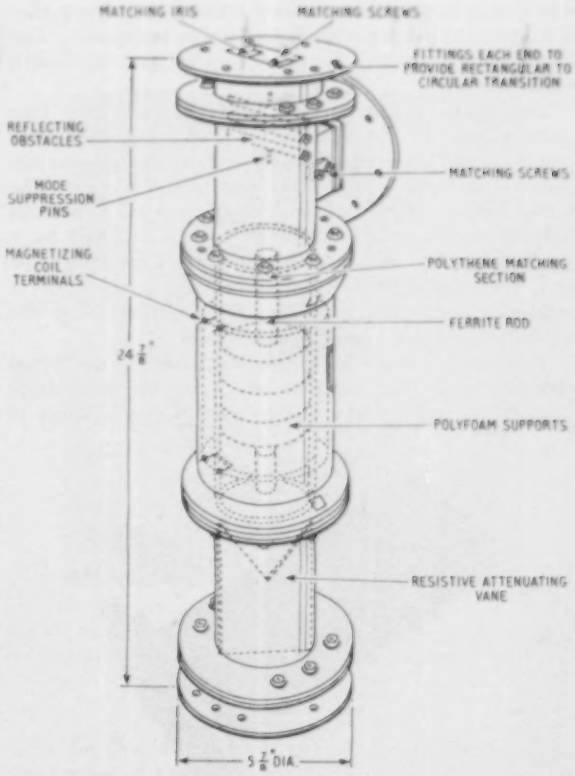

Fig. 3. *General construction of the type 1 ferrite switch*

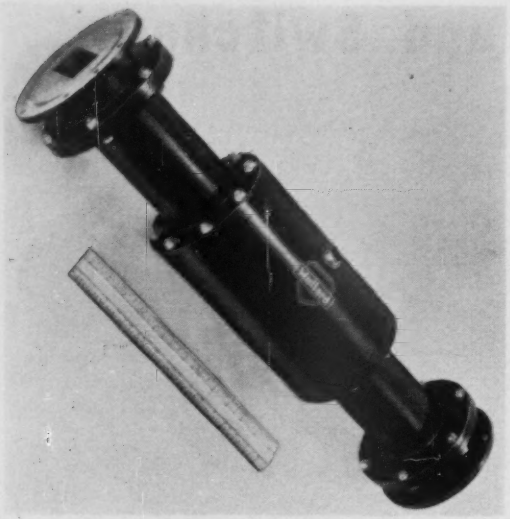

Fig. 4. *Ferrite switch, type 2, using 90° Faraday rotation*

reflection. In combination with a circulator the crystal assembly becomes a single-throw switch.

Ferrite Components

The material selected for the ferrite components was Ferroxcube 5B1, which has a sufficiently low value of saturation magnetization (about 1,000 Oe), to avoid the effect of low field loss. The diameter of the ferrite rod was chosen to give the necessary rotation in a few inches of material at magnetic fields well below saturation. The maximum diameter is, however, set by the bandwidth requirement.

The design of the ferrite components was such that each section of the component (i.e., the waveguide junctions and the ferrite-loaded section) was adequately matched independently of the rest of the system. The ferrite section was matched by means of a pair of polythene rods and the junctions were matched by a single inductive iris. The bandwidth of each junction was several-hundred megacycles per second for a reflection coefficient of 0·03. The bandwidth of the ferrite section was somewhat narrower.

The feature which limits the bandwidth of the ferrite components is the rotation characteristic and, while loading the waveguide reduces the rate of change of rotation with frequency, it is unnecessary for narrow-bandwidth components.

Static Circulator

This component, shown in Fig. 1, consists of a ferrite section, including an adjustable permanent-magnet system, and a pair of dual-mode transducers. Each component is independently matched to give a v.s.w.r. of >0·9, but set screws are included in each port so that very high peak isolations are obtainable between the physically adjacent ports. The performance of the circulator is as follows:

Insertion loss	< 0·2 dB
Isolation	> 30 dB
Centre frequency	3,000 Mc/s
Bandwidth	30 Mc/s
Input v.s.w.r.	> 0·9

Ferrite Switch (1)

This component (Fig. 2) may be used for connecting a pair of aerials alternately to a transmitter or to a receiver. The component is basically similar to the static circulator except that the third port is internally provided with a matched load (Fig. 3). The switching action is obtained by reversing the applied magnetic field in the ferrite section. The magnetic field is provided by a bifilar wound coil in which current flows in one half or the other, controlled by a pair of silicon controlled rectifiers.

In order to obtain rapid switching times the waveguide in the ferrite-loaded section is fabricated by thinly plating a tube of insulating material of the appropriate bore.

The performance of the switch is as follows:

Insertion loss	< 0·5 dB
Isolation	> 20 dB
Centre frequency	3,000 Mc/s
Bandwidth	100 Mc/s
Input v.s.w.r.	> 0·9
Switching time	< 150 μsec

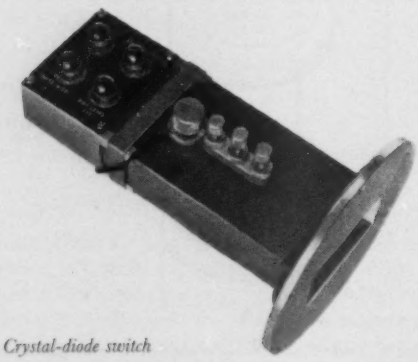

Fig. 5. *Crystal-diode switch*

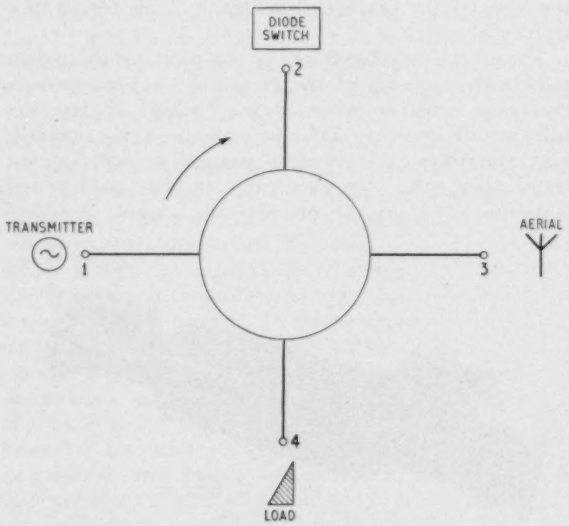

Fig. 6. *Connections of crystal switch and circulator*

Ferrite Switch (2)

This component (Fig. 4) is based on a 90° Faraday rotation and consists of a ferrite-loaded section and a pair of rectangular-to-cylindrical waveguide transducers in line. The magnetic field for the ferrite is provided by means of a coil wound outside the waveguide. The transducers include a matched load to absorb the cross-polarized component (this is similar to the one in Fig. 3). With zero field the plane of polarization of the incident energy is unchanged and it is transmitted through the component substantially unattenuated. With the current applied a 90° Faraday rotation is obtained and the energy is absorbed in the matched load provided. Any unabsorbed energy is reflected and reappears in the input waveguide as a small mismatch.

As with the previous switch, the ferrite-loaded section is fabricated by thinly plating a tube of insulating material of the appropriate bore.

Crystal Switch

A switched crystal (Fig. 5) may be used in combination with a circulator to form a non-reciprocal switch (Fig. 6). The crystal is matched under forward-bias conditions. Under back-biased conditions the crystal becomes substantially reactive and a reflection coefficient of 0·99 may be achieved. The crystal used was a CS3A which may only be operated with low r.f. powers (<1 mW) in the back-biased condition. The advantage of the crystal switch lies in the low video switching power required.

The performance of the crystal switch in conjunction with a circulator is as follows:

Insertion loss	< 0·5 dB
Isolation	> 30 dB
Centre frequency	3,000 Mc/s
Bandwidth	30 Mc/s
Input v.s.w.r.	> 0·9
Switching time	< 10 μsec

Printed-Circuit Plating

USE OF NOBLE METALS

By J. E. Philpott B.Sc.*

The electronics industry is relatively young, but it has already witnessed several fundamental developments, among which is the use of printed circuits. Originally, these circuits were designed to replace the wiring in conventional chassis, but it was soon recognized that they were capable of more specialized uses, of which some necessitated the application of noble metal electrodeposits. Circuits manufactured in this way have been applied to all types of equipment, ranging from domestic radio and television receivers to elaborate defence devices.

Several materials, and in consequence a variety of production techniques, were investigated in the initial stages of printed-circuit development and although some of these have proved useful for certain applications, most printed circuits now in use are based on copper foil bonded to a laminated phenolic resin. The complex electrical patterns of such circuits are produced by etching techniques.

In many applications all that is required of a printed circuit is that the tracks should serve to connect components and that a minimum of difficulty should be encountered in soldering operations.

Now that more extensive use is being made of printed circuits there is a wide field of application in which the contact properties of the track are of great importance, and this may be divided into two major categories.

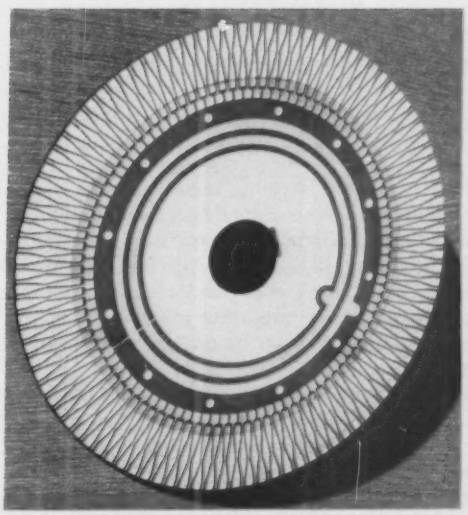

Fig. 1. *This nine-inch stator of a 200-way switch for telemetry equipment has electrodeposited rhodium on the contact surfaces of the 200 sectors and on the two slip rings at the centre*

In the first, the printed circuit is used as the major component of a switch, the track replacing the mechanically produced segments or slip rings of the conven-

* *Johnson, Matthey & Co., Limited.*

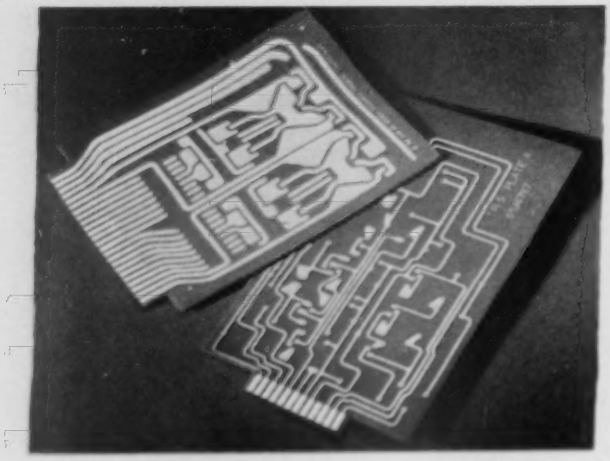

Fig. 2. *Two printed circuits by Ferranti Ltd. designed for plugging into fixed sockets. The contact surfaces are palladium plated*

tional switch, a suitable wiper being arranged to sweep across the contact surfaces (Fig. 1). In the second, one edge of the printed circuit board is designed for plugging into a suitably designed socket. This application is of particular importance to the computer manufacturer since it makes it possible for various sub-circuits, each capable of carrying out some mathematical operation, to be plugged into a computer during 'programming'. The plug contacts usually take the form of a number of parallel strips at right angles to one edge of the board (Fig. 2).

Track Materials

Considered purely as a contact material, copper is not ideal because it is liable to form a non-conducting oxide film and it is too soft for use in adverse mechanical conditions.

As a result, it is often necessary to electroplate the copper track with a noble metal in order to obtain a low, stable contact resistance or a long contact life. The choice of metal is determined by the conditions to which the printed circuit is to be subjected and usually rests between palladium, gold and rhodium although, in certain special cases, platinum may be chosen. All four metals remain free from oxidation at normal temperatures.

Electrodeposition of these metals on the printed circuit requires that a continuous connection, which may be only temporary, is made between all the sections to be plated.

With plug circuits the board can be extended by $\frac{1}{4}$ in. beyond the plugs, this extension having a band of copper linking the plugs together (Fig. 3). After plating, the extension is cut off. A similar procedure can sometimes be followed with switch circuits by extending the leads from the segments to a connecting strip. Where this is not practicable, the segments can often be extended to a ring in the centre of the switch, which is masked during plating. This ring can subsequently be etched away.

In the early days of electroplating printed circuits attack by the electrolyte on the printed circuit frequently occurred, and also highly-stressed deposits were produced. To minimize these effects it is essential to use a high-grade laminate produced specifically for printed circuits. Where doubt exists, prototype circuits should be experimentally plated before the main batch is made.

Palladium

Of the recommended metals, palladium shows the most useful combination of properties and also has the advantage of being the least costly. The bath from which it is deposited is initially neutral, although it becomes slightly more acid as it is worked, and the electrodeposit is not highly stressed. The 100% cathode efficiency of the process, with consequent freedom from hydrogen bubbles on the surface of the cathode, results in pin-hole free electrodeposits. The metal is therefore ideal for application to printed circuits.

Electrodeposited palladium is intermediate in hardness [300–400 V.P.N. (Vickers Pyramid Number)] between electrodeposited alloy gold and rhodium, and is of the same order of hardness as electrodeposited platinum. It is suitable for all plug applications, in conjunction with palladium plated or alloy-gold-plated sockets, and for many switch applications. A deposit in the thickness range 0.00005 to 0.0002 in. can be applied.

Platinum

The development of platinum plating has been limited owing to difficulties associated with depositing the metal from baths of a conventional type. Recently, however, a new bath based on the complex sulphato-dinitrito-platinous acid has been produced. Known as the DNS platinum-plating solution, it can be safely used for printed circuits. Electrodeposited platinum can be specified as an alternative to palladium, but its use in this field is largely unexplored.

Acid Hard Gold

As mentioned previously, in early attempts to electroplate printed circuits attack on the printed circuit

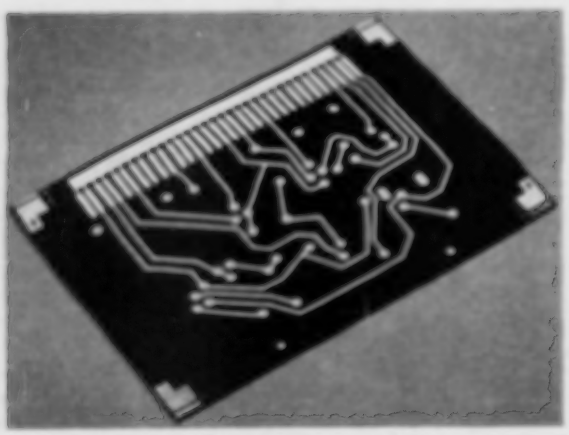

Fig. 3. *A terminal printed-circuit board after the end tags have been plated with palladium and before the connecting band has been cut off*

boards frequently occurred. This was often the case when deposits of gold were being laid down from baths of high cyanide content. In addition, fine gold, with a hardness of 65 V.P.N., was found to be too soft for many applications.

Recently, special plating baths have been developed giving alloy gold deposits with hardnesses of the order of 125 V.P.N., a value consistent with maximum durability. As implied by the name, the pH range of these baths is maintained in the acid region to minimize attack on printed circuits and to reduce the stress in the electrodeposits. Deposits in the thickness range 0.0001–0.0002 in. are recommended for plug applications, the mating contacts being plated with alloy gold or palladium, while for switch applications deposits in the thickness range 0.0005–0.001 in. are recommended.

TABLE 1
Electrodeposits for Printed Circuits

Deposit	Application	Hardness V.P.N.*	Electrical conductivity % I.A.C.S.†	Thickness of deposit inch	pH of bath
Palladium	Plugs or Switches	300 to 400	16	0·00005 to 0·0002	7·0
Alloy Gold	Plugs	120 to 130	70	0·0001 to 0·0002	6·0–7·0
Rhodium	Switches	800	35	0·00005 to 0·0002 direct or 0·000015 on a nickel undercoat	<1·0
Platinum	Plugs or Switches	300 to 400	15	0·00005 to 0·0001	0·5–2·0
Nickel	As undercoat only	200 to 300	29	0·00025 (undercoat for rhodium)	5·0–6·0

* Vickers Pyramid Number. † International Annealed Copper Standard

Rhodium

Rhodium is the hardest of the recommended metals with a value of 700–800 V.P.N., and its use on printed circuits should be confined to switch applications, a copper-silver-gold alloy being used for the wiper. The complete freedom of rhodium from tarnish films results in very low and stable values of contact resistance even under conditions of low contact pressure. Because the deposit usually applied is thin, seldom exceeding 0.0005 in., arcing or sparking in the switch will rapidly puncture the electrodeposit and result in early failure. Rhodium-plated contacts for switching duties should therefore be used only in circuits where both currents and voltages are very small.

Two different process specifications are available. Either a deposit 0.00005–0.0002 in. thick can be applied direct to the copper, or a nickel undercoat 0.00025 in. thick can be applied from a special nickel bath and the rhodium electrodeposit limited to 0.000015 in. The latter specification, which is widely used in the U.S.A., is steadily gaining favour in the U.K. since the higher internal stress inevitable with thicker rhodium deposits is avoided.

Table 1 shown here sets out the physical properties of the metals together with their major applications and serves as a guide in selecting a suitable metal for a given printed circuit.

TECHNICAL WRITING AWARDS

At a luncheon on 27th April the awards of 25-guinea premiums for six technical articles published during 1960 were presented by Sir Harold Bishop on behalf of the Electronic Engineering Association and the Radio Industry Council.

The prize-winners were:

C. M. Cade, M.Brit. I.R.E., S.M.I.R.E., M.Inst. Nav., "Indra-Red Radar Surveillance and Communications", *British Communications and Electronics*, June and July.

A. E. Crawford, M.Brit. I.R.E., S.M.I.R.E., "Piezo-Electric Voltage Transformers", *Wireless World*, October.

G. L. Grindale, B.Sc., Ph.D., and D. A. Paynter, A.M.I.E.E., A.M.Brit. I.R.E., "A Tropospheric Scatter Link over a 200-Mile Path", *Point-to-Point Telecommunications*, October.

P.L. Owen, M. F. Partridge, M.A. and T. R. H. Sizer, A.M.I.E.E., "The Differential Analyser and its Realisation in Digital Form", *Electronic Engineering*, October.

E. N. Rowlands and H. S. Wolff, "The Radio Pill", *British Communications and Electronics*, August.

D. L. Swale, "Integrated Electromechanical Design as Applied to Electronic Equipment", *British Communications and Electronics*, March.

It was announced at the luncheon that for 1961 there will be eight premiums of 30 guineas each and that awards will be made for purely technical articles as well as for those of commercial or interpretive merit.

BRIT.I.R.E. "RADIO TECHNIQUES AND SPACE RESEARCH" CONVENTION

The British Institution of Radio Engineers 1961 Convention "Radio Techniques and Space Research" is to be held in Christ Church, Oxford, from 5th to 9th July.

The programme will include sessions on Satellite Engineering, Navigation Satellites, Communication Satellites, Instruments in Satellites, Extra-Terrestrial Measurements, Satellites and Launcher Control, Telemetry and Radio Astronomy.

Further information and registration forms may be obtained from the Institution offices at 9 Bedford Square, London, W.C.1.

Bode's Variable Equalizer

DESIGN CRITERIA

By S. S. Hakim, Ph.D., B.Sc.*

Very often the need arises for the design of an equalizer in which the insertion-loss characteristic can be varied so as to compensate, for example, for the variation of the loss of a cable with temperature. Bode[1] has described several types of variable equalizers in which the insertion-loss characteristic can be varied by the variation in the value of one of its elements. One such element can be a thermistor. In this article, the basic theory is developed for one particular type of Bode's variable equalizer that is most versatile. This is followed by a typical design example.

Basic Theory

Consider Fig. 1 showing one type of Bode's variable equalizer connected between a load resistance R and a voltage source of resistance also equal to R. The symmetrical four-pole network is assumed to have an

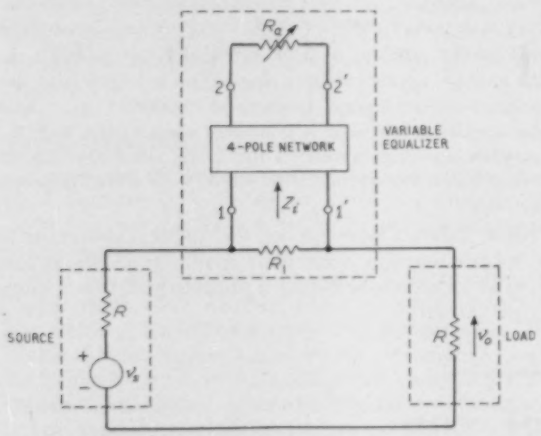

Fig. 1. *One type of Bode's variable equalizer*

image impedance of R_a and an image propagation constant of θ. The resistor R_a is the variable element, and the significance of the resistor R_1 will be pointed out later.

The driving point impedance Z_i measured at the terminal pair 11′ of the four-pole network, with terminals 22′ terminated with R_a, is given in terms of its image impedance R_0 and propagation constant θ by Seeley and Le-Page[2]

$$Z_i = R_0 \frac{R_0 (e^{2\theta} - 1) + R_a (e^{2\theta} + 1)}{R_0 (e^{2\theta} + 1) + R_a (e^{2\theta} - 1)} \quad .. \quad .. \quad (1)$$

Let ρ be the reflection coefficient, with respect to R_a, at the terminal pair 22′ when terminals 11′ of the four-pole network are closed with resistor R_0, then

$$\rho = \frac{R_a - R_0}{R_a + R_0} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

from which one can deduce that

$$\frac{R_a}{R_0} = \frac{1 + \rho}{1 - \rho} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

Combining Equs (1) and (3) gives

$$Z_i = R_0 f(\theta) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (4)$$

Where

$$f(\theta) = \frac{(e^{2\theta} - 1)(1 - \rho) + (e^{2\theta} + 1)(1 + \rho)}{(e^{2\theta} - 1)(1 + \rho) + (e^{2\theta} + 1)(1 - \rho)} = \frac{e^{2\theta} + \rho}{e^{2\theta} - \rho} \quad (5)$$

From Fig. 1, the voltage v_0 developed across the load is given by

$$v_o = v_s \cdot \frac{R [R_1 + R_0 f(\theta)]}{2RR_1 + f(\theta) R_0 (2R + R_1)} \quad .. \quad .. \quad (6)$$

However, when the variable equalizer is removed as in Fig. 2, then the voltage v_0' developed across the load is noted to be

$$v_o' = \frac{v_s}{2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (7)$$

Hence, the insertion voltage ratio of the variable equalizer is deduced from Equs (6) and (7) to be

$$\frac{v_o'}{v_o} = e^{\phi} = 1 + \frac{R_1 R_0 f(\theta)}{2R [R_1 + R_0 . f(\theta)]} \quad .. \quad .. \quad (8)$$

If the resistor R_1 is chosen so as to satisfy the condition

$$R_1^2 = R_0^2 \left(1 + \frac{R_1}{2R}\right) \quad .. \quad .. \quad .. \quad .. \quad (9)$$

that is,

$$R = \frac{R_1 R_0^2}{2 (R_1^2 - R_0^2)} \quad .. \quad .. \quad .. \quad .. \quad (10)$$

then eliminating R between Equs (8) and (10) gives:

$$e^{\phi} = \frac{R_1}{R_0} \frac{1 + \frac{R_1}{R_0} f(\theta)}{\frac{R_1}{R_0} + f(\theta)} \quad .. \quad .. \quad .. \quad .. \quad (11)$$

Let $\phi = \alpha_0$ when the variable resistor R_a has its mid-setting value of R_0; then Equs (2) and (4) give,

* *The General Electric Co. Ltd., Coventry.*

respectively, $[\rho]_{R_a=R_0} = 0$ and $[f(\theta)]_{R_a=R_0} = 1$ in which case Equ. (11) reduces to

$$e^{\alpha_0} = \frac{R_1}{R_0} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (12)$$

Eliminating R_1/R_0 between Equs (11) and (12) gives

$$e^{\phi-\alpha_0} = \frac{1 + e^{\alpha_0} f(\theta)}{e^{\alpha_0} + f(\theta)} \quad .. \quad .. \quad .. \quad .. \quad (13)$$

from which one can deduce that

$$\frac{e^{\phi-\alpha_0} - 1}{e^{\phi-\alpha_0} + 1} = \tanh\left(\frac{\phi - \alpha_0}{2}\right) = \frac{K}{2}\frac{f(\theta) - 1}{f(\theta) + 1} \quad .. \quad (14)$$

where

$$\frac{K}{2} = \frac{e^{\alpha_0} - 1}{e^{\alpha_0} + 1} = \tanh\left(\frac{\alpha_0}{2}\right) \quad .. \quad .. \quad .. \quad (15)$$

Using Equs (5) and (14) gives

$$\tanh\left(\frac{\phi - \alpha_0}{2}\right) = \frac{K}{2}\rho e^{-2\theta} \quad .. \quad .. \quad .. \quad (16)$$

that is

$$\begin{aligned}\phi - \alpha_0 &= 2\tanh^{-1}\left(\frac{K}{2} \cdot \rho e^{-2\theta}\right) \\ &= K\rho e^{-2\theta} + \frac{1}{12}(K\rho e^{-2\theta})^3 \\ &\quad + \frac{1}{80}(K\rho e^{-2\theta})^5 + \ldots . \quad .. \quad .. \quad (17)\end{aligned}$$

In practice, the term $K\rho e^{-2\theta}$ is sufficiently small to justify neglecting the second and higher terms in the series expansion of Equ. (17) and so

$$\phi - \alpha_0 \simeq K\rho \cdot e^{-2\theta} \quad .. \quad .. \quad .. \quad .. \quad (18)$$

From Equ. (2) one notes that $[\rho]_{R_a=0} = -1$ and $[\rho]_{R_a=\infty} = 1$ and so from Equ. (18) it is deduced that the insertion-loss characteristic corresponding to $R_a = 0$

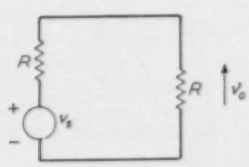

Fig. 2. *Fig. 1, with the variable equalizer removed*

is the image, with respect to the mid-setting loss (i.e., when $R_a = R_0$), of the characteristic when $R_a = \infty$.

It is to be noted that this 'symmetry' property of the insertion-loss characteristic of the variable equalizer of Fig. 1 is achieved only when Equ. (9) is satisfied.

Let $\phi = \alpha + j\beta$

$$\theta = A + j(B + \psi) \quad .. \quad .. \quad .. \quad .. \quad (19)$$

then Equs (18) and (19) give

$$\begin{aligned}(\alpha - \alpha_0) + j\beta &\simeq K\rho e^{-2A-2j(B+\psi)} \\ &= K\rho e^{-2A}\{\cos 2(B+\psi) - j\sin 2(B+\psi)\}\end{aligned}$$

Where, on equating the real and imaginary terms, one obtains

$$\alpha - \alpha_0 = K\rho e^{-2A} \cdot \cos 2(B + \psi) \quad .. \quad .. \quad (20)$$

$$\beta = -K\rho e^{-2A} \sin 2(B + \psi) \quad .. \quad .. \quad .. \quad (21)$$

Solving Equs (20) and (21) for A and $B + \psi$ gives

$$\begin{aligned}A &= -\tfrac{1}{4}\log_e\left\{\left(\frac{\alpha - \alpha_0}{K\rho}\right)^2 + \left(\frac{\beta}{K\rho}\right)^2\right\} \\ &= -\tfrac{1}{4}\log_e(a^2 + b^2) \quad .. \quad .. \quad .. \quad (22)\end{aligned}$$

$$\begin{aligned}B + \psi &= -\tfrac{1}{2} \cdot \tan^{-1}\left(\frac{\beta}{\alpha - \alpha_0}\right) \quad .. \quad .. \quad (23) \\ &= -\tfrac{1}{2} \cdot \tan^{-1}\left(\frac{b}{a}\right) \quad .. \quad .. \quad .. \quad (23)\end{aligned}$$

where

$$\left.\begin{aligned}a &= \frac{\alpha - \alpha_0}{K\rho} \\ b &= \frac{\beta}{K\rho}\end{aligned}\right| \quad .. \quad .. \quad .. \quad .. \quad .. \quad (24)$$

Equs (10), (12), (22), (23) and (24) are the necessary design equations. Normally α_0 is given and $\alpha - \alpha_0$ is specified as a function of frequency for a known value of R_a and it is required to determine R, R_1 and the 4-pole network.

Now the insertion voltage ratio of any 2-pole network is necessarily a minimum-phase one. Hence, the insertion loss α and phase-angle β are uniquely related to each other, and so if α is given then β can be determined from it by known methods (Bode[3]). Knowing α, β and $K\rho$, Equs (22), (23) and (24) can be used to determine the image loss A and phase-angle $B + \psi$ of the 4-pole network. The loss A is next realized by a constant-resistance bridged-T type of fixed equalizer. The phase-angle B associated with this fixed equalizer can, therefore, be determined. Finally, the residual phase-angle ψ is deduced and realized by a constant-resistance type of phase equalizer (i.e., an all-pass structure) and the design is complete.

It is, however, to be noted that for the 4-pole network to be physically realizable, it is necessary that at all frequencies $(\alpha - \alpha_0)/K\rho$ must be such that Equ. (22) always yields a positive value for A.

Distortion

Neglecting the second and higher terms in the series, expansion of Equ. (17) leads to a small error D between the desired characteristic and that realized according to Equ. (18). Consider the effect of neglecting the second term in the series expansion of Equ. (17)

$$\begin{aligned}\frac{1}{12} \cdot (K\rho e^{-2\theta})^3 &= \frac{K^3\rho^3}{12} \cdot \{e^{-2A-j2(B+\psi)}\}^3 \\ &= \frac{K^3\rho^3}{12} \cdot \{e^{-2A} \cdot \cos 2(B+\psi) - j\sin 2(B+\psi)\}^3\end{aligned}$$

the real part of which can be deduced to be

$$\begin{aligned}&\frac{K^3\rho^3}{12} e^{-6A} \cos 2(B+\psi)\{\cos^2 2(B+\psi) - 3\sin^2 2(B+\psi)\} \\ &= \frac{K^3\rho^3}{12} e^{-6A} \cos 2(B+\psi)\{4\cos^2 2(B+\psi) - 3\} \\ &\qquad .. \quad .. \quad (25)\end{aligned}$$

which is effectively equal to the error D.

Design Example

Let the design requirements be $\alpha_0 = 6$ dB $= 0{\cdot}69$ nepe*r* and the $(\alpha - \alpha_0)$ characteristic corresponding to

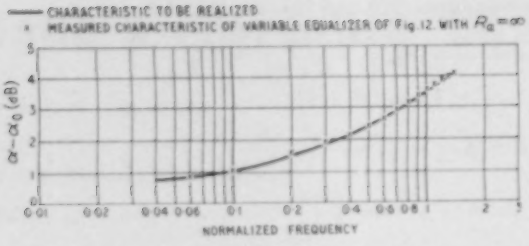
CHARACTERISTIC TO BE REALIZED
MEASURED CHARACTERISTIC OF VARIABLE EQUALIZER OF Fig. 12. WITH $R_0 = \infty$
$\alpha - \alpha_0$ (dB)
NORMALIZED FREQUENCY

Fig. 3. $\alpha - \alpha_0$ *plotted as a function of the normalized frequency* ω

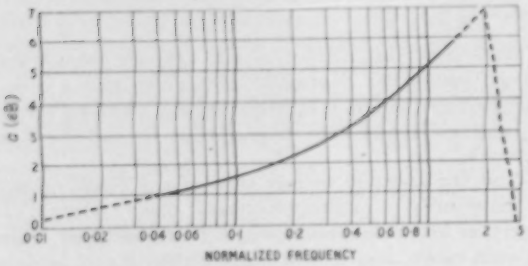

Fig. 4. *Loss characteristic 'a' plotted against normalized frequency* ω

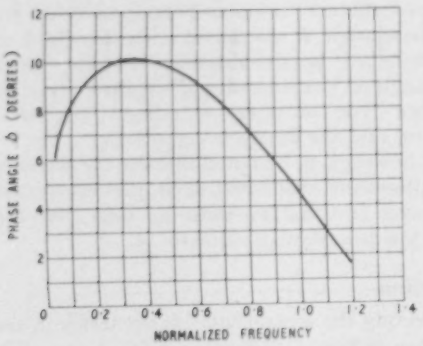

Fig. 5. *Phase-angle 'b' plotted against normalized frequency* ω

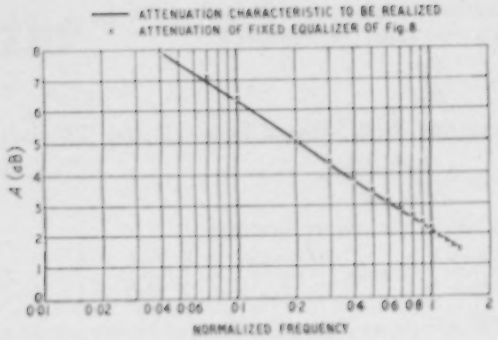

Fig. 6. *Loss 'A' plotted against normalized frequency* ω

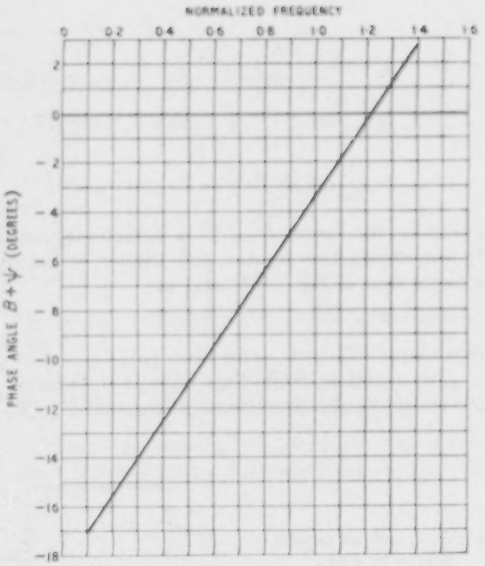

Fig. 7. *Phase-angle* $B + \psi$ *plotted against normalized frequency* ω

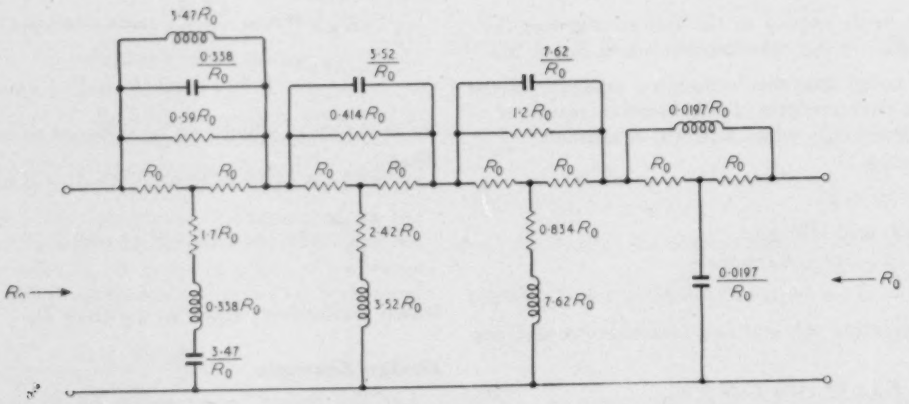

Fig. 8. *Fixed equalizer*

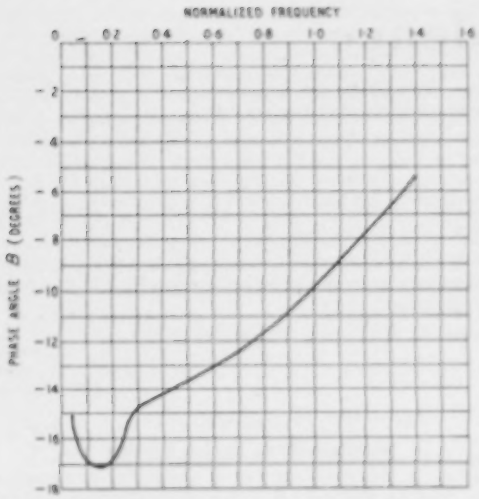

Fig. 9. *Phase-angle 'B' plotted against normalized frequency* ω

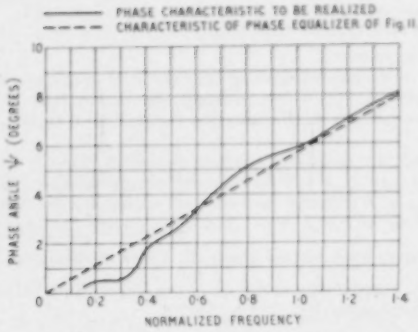

Fig. 10. *Phase-angle* ϕ *plotted against normalized frequency* ω

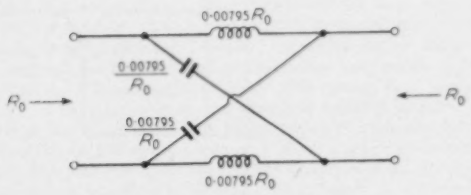

Fig. 11. *Phase equalizer*

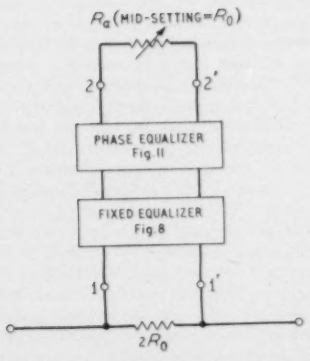

Fig. 12.
Designed variable equalizer

$R_a = \infty$ (i.e., $\rho = 1$) be as shown plotted in Fig. 3 as a function of the normalized frequency ω.

Thus Equ. (12) gives $R_1 = 2R_0$

Equ. (10) gives $R = \frac{R_0}{3}$

Equ. (18) gives $K = \frac{2}{3}$

and so $[K\rho]\ R_{a=\infty} = \frac{2}{3}$.

Hence, using Equ. (24) and the characteristic of Fig. 3 one can determine the loss characteristic (a) which is shown plotted in Fig. 4. Since the frequency range of interest is $\omega = 0{\cdot}1$ to $\omega = 1{\cdot}4$ then one can complete the characteristic of Fig. 4 outside the frequency band as shown dotted. It may be found necessary to modify the dotted characteristic of Fig. 4 several times in order to ensure that the residual phase ϕ is realizable.

The loss characteristic of Fig. 4 is then approximated by a plurality of straight lines and the associated phase (b) is then determined using the tables of Thomas[4], and the resulting phase characteristic is as shown in Fig. 5.

Using Figs. 4 and 5 and Equs (22) and (23) one can calculate A and $B + \phi$ which are shown plotted in Figs. 6 and 7, respectively. The loss A of Fig. 6 is then realized by the constant-resistance bridged-T equalizer of Fig. 8 whose actual loss is also indicated in Fig. 6. The phase angle B associated with the fixed equalizer of Fig. 8 is determined and is shown plotted in Fig. 9. Using Figs. 7 and 9 one can deduce the residual phase ϕ which is shown plotted in Fig. 10 and this is then approximated by the dotted straight line which can be realized by the all-pass structure of Fig. 11.

The completed design configuration is shown in Fig. 12 and the insertion-loss characteristic, corresponding to $R_a = \infty$, is also shown in Fig. 3, where the closeness of the actual and desired characteristics is noted to be sufficiently accurate.

Acknowledgements

The author is most grateful to Mr. D. W. Robson for useful discussion and comments; to Miss P. Bennett for building the variable equalizer network and obtaining the experimental measurements, and to The General Electric Co. Ltd. for permission to publish this paper.

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[2] W. R. Le Page and S. Seely, "General Network Analysis", McGraw-Hill, 1952.
[3] H. W. Bode, "Network Analysis and Feedback Amplifier Design", New York, Van Nostrand, 1945.
[4] D. E. Thomas, "Tables of Phase Associated with a Semi-Infinite Unit Slope of Attenuation", *Bell Syst. tech. J.*, October 1947.

LABORATORY APPARATUS AND MATERIALS EXHIBITION

Seventy leading British and overseas firms manufacturing laboratory instruments and materials will participate in the Laboratory Apparatus and Materials Exhibition which is being held from 19th—22nd June at the Royal Horticultural Society's New Hall, Westminster, London.

Twelve lecturers will speak in the R.H.S. lecture hall during the exhibition, which is being organized by U.T.P. Exhibitions Ltd., 9 Gough Square, London, E.C.4.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Zener Diode in Stabilized Transistor Power Supplies

SIR,—Though a stabilizing element (e.g., a Zener diode) is not needed to deliver appreciable current when used in a stabilized power supply of the valve type, this is not generally true when transistors are employed. That a certain inequality is to be satisfied between the maximum permissible Zener diode current, the maximum load current in shunt with the Zener diode and the

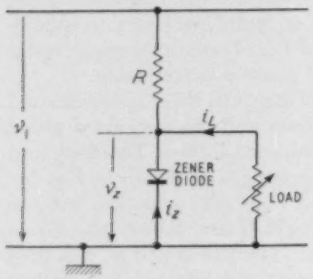


Fig. 1

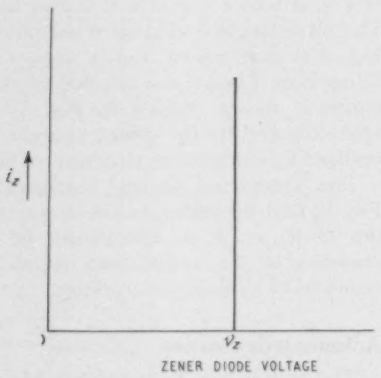


Fig. 2

percentage input voltage variation before the system becomes practically realizable is evident from the following considerations.

Consider the circuit in Fig. 1 with the following notations.

i_L = load current
i_z = Zener diode current
v_i = input voltage
v_z = Zener diode voltage

The arrowheads indicate the positive directions of current. Now, suppose that the maximum load current = $i_{L\alpha}$ when the applied voltage falls to a minimum given by $v_{i\beta}$. Then Zener diode current is also a minimum = $i_{z\beta}$. Thus from Fig. 1

$$\frac{v_{i\beta} - v_z}{R} = i_{z\beta} + i_{L\alpha} = i_1 \quad \text{(say)} \qquad \ldots \quad \ldots \quad \ldots \quad (1)$$

Again suppose that the minimum load current = $i_{L\beta}$ when the applied voltage rises to a maximum value given by $v_{i\alpha}$, then Zener diode current has a maximum value = $i_{z\alpha}$. Thus as in Equ. (1),

$$\frac{v_{i\alpha} - v_z}{R} = i_{z\alpha} + i_{L\beta} = i_2 \quad \text{(say)} \qquad \ldots \quad \ldots \quad \ldots \quad (2)$$

From (1) and (2), $$R = \frac{(v_{i\alpha} - v_{i\beta})}{(i_2 - i_1)} \qquad \ldots \quad \ldots \quad \ldots \quad (3)$$

and from (1) and (3), $$v_{i\beta} = \frac{(v_{i\alpha} - v_{i\beta})}{(i_2 - i_1)} i_1 + v_z \qquad \ldots \quad \ldots \quad (4)$$

Considering an idealized Zener characteristic as shown in Fig. 2, for which $i_{z\beta} = 0$, then from (1),

$$i_1 = i_{z\beta} + i_{L\alpha} = i_{L\alpha} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

Also supposing that the minimum load current $i_{L\beta} = 0$, then from (2),

$$i_2 = i_{z\alpha} + i_{L\beta} = i_{z\alpha} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

Equations (3), (4), (5) and (6) together give,

$$R = \frac{v_{i\alpha} - v_{i\beta}}{i_{z\alpha} - i_{L\alpha}} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

and $$v_{i\beta} = \frac{v_{i\alpha} - v_{i\beta}}{i_{z\alpha} - i_{L\alpha}} \cdot i_{L\alpha} + v_z \qquad \ldots \quad \ldots \quad \ldots \quad (8)$$

Writing $v_{i\alpha} - v_{i\beta} = \lambda v_{i\beta}$, $\lambda > 0$ in Equ. (8)

$$v_{i\beta} = \frac{v_z (i_{z\alpha} - i_{L\alpha})}{i_{z\alpha} - (1 + \lambda) i_{L\alpha}} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

In a practically realizable case, R must be positive. Therefore from (7), since $(v_{i\alpha} - v_{i\beta})$ is necessarily positive, by definition

$$i_{z\alpha} > i_{L\alpha} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

Again applying the condition of realizability of R [i.e., Equ. (10)] to (9),

$$i_{z\alpha} > (1 + \lambda) i_{L\alpha} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

since $v_{i\beta}$ is positive.

Thus the necessary condition for the case to be practically realizable is given by Equ. (11) which automatically satisfies Equ. (10).

We now proceed to illustrate the method by a simple example as follows.

Consider the stabilized circuit shown in Fig. 3. Suppose $v_{i\alpha} = 1{\cdot}5\, v_{i\beta}$ (i.e., $\lambda = 0{\cdot}5$) the maximum permissible Zener diode current $i_{z\alpha} = 8$ mA and that we want to stabilize for a maximum current drain of $I = 2$ A (= 2000 mA). Then referring to Fig. 3 the maximum load current across the Zener diode $i_{L\alpha} = \dfrac{2000}{\beta}$ mA where β = current gain of the series transistor. Equ. (11) demands that for stabilization $8 > 1{\cdot}5 \times 2000/\beta$

or, $\beta > 375$ $\qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$

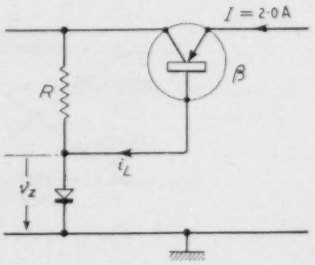


Fig. 3

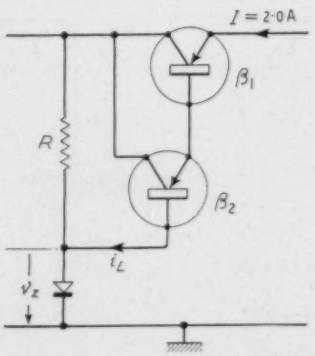


Fig. 4

which is rather unlikely to be obtained at 2 A. However, assuming $\beta = 30$, which is more feasible, the circuit could be realized if $i_{z\alpha} > 1{\cdot}5 \times 2000/30$ or, $i_{z\alpha} > 100$ mA, which is also rather unusual for a small Zener diode. Alternatively, condition (12) could easily be achieved with two transistors as shown in Fig. 4 where the effective $\beta = \beta_1 \beta_2$. Assuming a β of the order of 30 for each, $\beta = 900$ which satisfies (12). Equ. (11), however, shows that for a maximum Zener diode current of 8 mA and $\lambda = 0{\cdot}5$ (as before), $i_{L\alpha} < 5{\cdot}3$ mA so that for a β of 30, the maximum current drain permissible $= (30 \times 5{\cdot}3)$ or 159 mA with the circuit configuration of Fig. 3.

In conclusion, we would like to take this opportunity of thanking the directors of Pye Ltd. for their kind permission to publish this note.

R. R. Dutta Gupta.
B. Tyler.

Pye Limited,
Cambridge.
28th March 1961.

On the Action of Transistors

Sir,—Some recent remarks in this journal[1,2,3] are all favourable to the viewpoint that the connection of a junction transistor so that the emitter is the common terminal is the important one, and should be considered the standard. The present purpose is to urge that it would help yet further to consider as the important parameters of the transistor the transconductance[4] and the input admittance. This amounts to taking as the practical equivalent circuit that shown in Fig. 1. The 'equivalent' capacitance at the input terminal (i.e., the base) will usually need to be taken into account, but the trans-

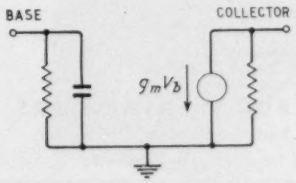


Fig. 1

conductance can, to an approximation good enough for many purposes, be taken as constant, independent of frequency[4]. Then the analysis of the circuit becomes very similar to that of circuits using vacuum tubes.

In the equivalent circuit there should strictly be a capacitance shunting the output. This can be added if it should seem necessary: if transistors are cascaded, however, the much larger 'equivalent' capacitance at the input of the next transistor will usually dominate the situation. The 'feedback' from collector to base within the transistor is also not shown; in practical work it is often either ignored or else compensated by things which are adjusted by trial.

It seems to be felt that a viewpoint which makes the transistor so very like the vacuum tube is not quite proper, on the ground that the transistor is 'operated by current, not voltage'. It is desired here to challenge that opinion. By 'current' one means 'charge in motion'; thus, if a device is 'operated by current', its operation must depend on the motion, and not just the presence, of the charges; in other words it must be a measure of its velocity. There is not an iota of reason for believing that a transistor is 'controlled by current' in that sense. Rather, is it not like the arrangement of a parallel-plate capacitor shunted by a resistor? The force between the plates is proportional to the current supplied to the combination, but no one would say that the force is 'controlled by current, not voltage'.

Once this point be granted, the way is clear, incidentally, to see an interesting parallel between the transistor and the vacuum triode. This is illustrated in Fig. 2. After an 'emitting region' is an 'insulating region permeable to carriers': being just vacuum in the tube and the depletion region between the emitter and base in the transistor. After that comes a 'conducting region permeable to carriers': viz., the grid of the tube, and the base region of the transistor. That is followed by another 'insulating region permeable to carriers': being again vacuum in the tube and, in the transistor, the depletion region between the base and the collector. Finally there is a 'collecting region', which is the anode of the tube and the collector of the

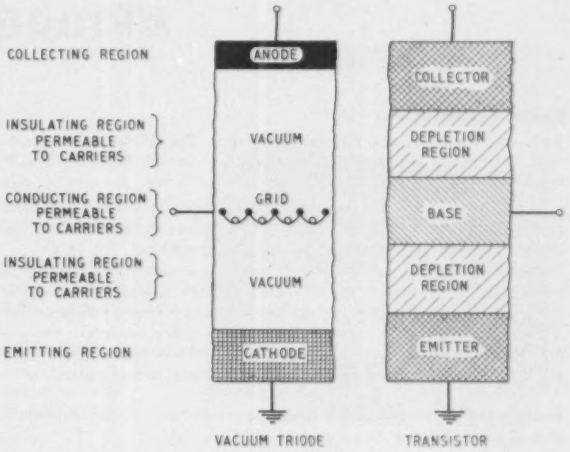


Fig. 2

transistor. Thus the two structures are entirely analogous; they differ essentially only in that (a) the base of the transistor is a much more efficient shield between collector and emitter than is the grid of the tube between anode and cathode, thus the base must be run in that polarity in which it will help the emission of carriers from the emitter, and (b) the depletion regions in the transistor are poorer insulators than is the vacuum in the tube.

This way of looking at the similarity of the two devices may be especially useful for purposes of instruction, in showing that, to a considerable extent, the same way of thinking may be applied to both.

H. L. Armstrong.

Department of Physics,
Queen's University,
Kingston, Ontario, Canada.
21st April 1961.

REFERENCES

[1] Editorial, *Electronic Technology*, March 1961, Vol. 38, p. 75.
[2] R. W. Jackson, ibid., p. 87
[3] B. Chatterjee, ibid., p. 108.
[4] H. L. Armstrong, *Proc. Inst. Radio Engrs*, Vol. 47, p. 83, 1959.

SATELLITE AND COMMUNICATIONS

Communication Satellites Inc., a subsidiary of the U.S. General Electric Co., has applied for authority to establish a communication network embodying satellites. The plan is to place 10 satellites equally spaced around the earth in an equitorial orbit and at about 6,000 nautical miles from the earth. It is estimated that they will remain in orbit for about 500 years but the useful life for communications is unlikely to exceed five years.

Each satellite will be about 25 ft in diameter and weigh some 60 lb. It will carry a receiver and transmitter powered by solar batteries and so act as a relay station. Transmissions to the satellite will be on 6,000 Mc/s and from it on 4,000 Mc/s.

The plan is to have 1,200 high-quality voice-frequency channels using f.m. and one television channel but it is expected to be able to add a further 300 voice-frequency channels of lower quality at a later date.

The satellites will not be 'stationary' viewed from the earth, but will have a time of orbit of 7·4 hours apparent. The real time will be 6·33 hours.

It is estimated that the cost of each satellite will be $3.5M and that the launching will cost $8M. The total cost of the project will be around $250M.

R.E.C.M.F. ELECTIONS

Mr. Arthur Bulgin (A. F. Bulgin & Co. Ltd.) has been elected president of the Radio and Electronic Component Manufacturers' Federation, in succession to Mr. E. M. Lee (Belling & Lee Ltd.).

Dr. G. A. V. Sowter (Telcon Metals Ltd.) succeeds Mr. Hector V. Slade (Garrard Engineering & Manufacturing Co. Ltd.) as chairman.

New Books

Radio Transmitters

By Laurence Gray and Richard Graham. Pp. 462 + xi. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 97s.

" . . . no attempt has been made to go into the elementary principles of electricity. Conversely, the book does not give mathematical derivations of the equations presented throughout the text. . . . Rather the book is intended to present a practical analysis of transmitter operation in all its various phases, together with usable information pertaining to specific problems in transmitter operation. Thus for those whose problems demand a more complete analysis or additional mathematical support, the reference lists provided at the end of each chapter will indicate additional source material."

Principles of Semiconductors: Introduction to the Properties and Applications of Semiconducting Materials

By M. G. Scroggie, B.Sc., M.I.E.E. Pp. 156 + viii. Published first in the U.S.A. by the Gernsback Library Inc. under the title "Fundamentals of Semiconductors" and now in a revised form for *Wireless World* by Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1. Price 21s.

"This book surveys the whole field of semiconductors and their numerous and diverse applications. Although first published in America, it has been completely anglicized and re-illustrated. . . .

One of the most important aims of this book is to provide sufficient basic theory for the reader, new to the subject, to easily understand the more advanced literature and articles regularly published in technical journals. The second aim is to explain the special properties of semiconductors and illustrate how these are applied to many kinds of useful devices."

Wires and R.F. Cables

By G. W. A. Dummer, M.B.E., M.I.E.E., and W. T. Blackband, M.Sc., A.M.I.E.E. Pp. 240 + xiii. Sir Isaac Pitman & Sons Ltd., Parker Street, London, W.C.2. Price 47s. 6d.

"This book has been written from the viewpoint of the user, to enable him to choose the best component for his particular requirements and to understand its fundamental characteristics. Comparison charts have been prepared . . . Future developments are covered . . .".

Thyratrons

By C. M. Swenne. Pp. 75. Published by Philips Technical Library. Price 12s. 6d.

"This booklet is intended to be an introduction to a small but significant field of industrial electronics: the application of thyratrons. It is specially addressed to the engineer, who is, on account of his profession, already familiar with electrical engineering, but who is more and more confronted with electronics."

Industrial Electronics Apparatus

By P. Van Der Ploeg. Pp. 98 + ix. Published by Philips Technical Library. Price 9s. 6d.

"The object of this booklet is to show both the designer and the service engineer the extent to which trouble-free operation of electronic equipment is dependent upon the little things that matter in its design, production, use and maintenance."

An Introduction to the Cathode Ray Oscilloscope. 2nd Edn. By Harley Carter, A.M.I.E.E. Pp. 123 + x. Published by Philips Technical Library. Price 15s.

Hi-Fi Amplifier Circuits

By E. Rodenhuis. Pp. 105 + ix. Published by Philips Technical Library. Price 15s.

The above four booklets are available from Cleaver-Hume Press Ltd., 31 Wright's Lane, Kensington, London, W.8.

Initiation a L'Electronique

By R. Faure. Pp. 356 + xviii. Published in French by Dunod Editeur, 92 rue Bonaparte, Paris (6e). Price 29 NF.

Zener Diode Handbook

Pp. 96. Published by International Rectifier Corporation, El Segundo, California, U.S.A. Price $2.00.

"A study of zener diodes and zener reference elements is prefaced by an introduction to semiconductor theory and reverse breakdown, and followed with a detailed and illustrated examination of design considerations, thermal and operating characteristics, and discussions of zener diode applications."

B. Van der Pol: Selected Scientific Papers (Vols. 1 and 2)

Edited by H. Bremmer and C. J. Bouwkamp. Published with the support of N. V. Philips' Gloeilampenfabrieken by North Holland Publishing Company, Amsterdam, P.O. Box 103. Price for Vols. 1 and 2 £6 13s. 0d.

"By these two volumes we attempt to cover the total scientific work of Balthasar van der Pol—in so far as this work is laid down in his published papers."

Measurement of Total Acoustic Power of Sources of Sound in a Reverberation Chamber

By K. A. Morch. Pp. 25. Published by the Acta Polytechnica Scandinavica Publishing Office, Box 5073, Stockholm 5, Sweden. Price Sw. Kr. 7.00.

"This report summarizes the theory of the reverberant sound field, and an account of the reverberation chamber investigations at the Laboratory of Technical Physics (Technical University of Denmark) is given. Investigations have been carried out at frequencies between 2 and 15 kc/s."

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for April 1961*

Date 1961 April	MSF 60 kc/s 1500 U.T. Parts in 10^9	Droitwich 200 kc/s 1030 U.T. Parts in 10^9
1	− 15·2	+ 4
2	− 15·3	+ 4
3	− 15·4	+ 7
4	− 15·5	+ 8
5	− 15·9	+ 7
6	− 15·3	+ 8
7	− 14·6	+ 8
8	− 14·4	+ 8
9	− 14·8	+ 7
10	− 15·1	+ 7
11	− 14·6	+ 7
12	NM	+ 8
13	− 15·2	+ 5
14	− 15·3	+ 6
15	− 14·9	NM
16	− 14·8	NM
17	− 14·7	+ 7
18	− 14·2	NM
19	− 14·9	+ 10
20	− 14·9	+ 9
21	− 15·3	+ 10
22	− 15·8	+ 11
23	− 15·9	+ 10
24	− 14·0	+ 10
25	− 14·0	+ 11
26	− 14·4	+ 12
27	− 14·3	+ 12
28	NM	+ 12
29	− 14·1	+ 12
30	− 14·0	+ 12

* *Nominal frequency corresponds to a value of* 9 192 631 770 *c/s for the caesium* $F,m(4,0) - F,m(3,0)$ *transition at zero field. NM = Not Measured.*

New Products

Square-Wave Generator

A square-wave unit manufactured by Feedback Ltd. has been designed to provide a square-wave driving signal.

The unit, type SW98 generates square waves over the frequency range 50 c/s to 100 kc/s when driven from an external sine-wave signal source; it also generates a square-wave signal at mains frequency. The rise-time is about 2 μsec for any operating frequency and a mark-space ratio trim is provided to enable an exact 1 : 1 ratio to be obtained.

The minimum sine-wave input signal amplitude for satisfactory operation is 2 V peak. The output is continuously variable from 0—50 V peak-to-peak independently of the input signal amplitude.

In addition, a circuit has been incorporated to generate a 2·5-kc/s calibration signal which may be switched to occur during either the positive or negative half of the square wave.

Feedback Ltd.,
Crowborough, Sussex.

Latching Relays

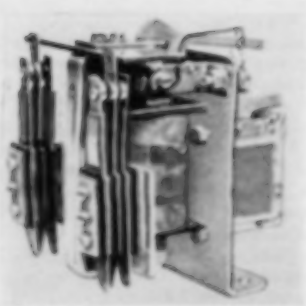
Elremco have recently extended their range of relays by the introduction of two latching-type units UNIV/LA and UNIV/LA/S.

The latching relay type UNIV/LA can carry up to eight contact sets, normally open, normally closed, changeover or mixed. The relay is held by a simple latching mechanism in the energized position after a short pulse is applied to the coil. When the lever attached to the relay is pressed, the relay contacts are released. The range of coils available covers requirements up to 440 V a.c. or 230 V d.c.

The latching relay type UNIV/LA/S is similar to the hand reset version (UNIV/LA) but can be both latched and unlatched by and electrical pulse.

Maximum contact ratings for both models are 6 A at 230 V a.c., 2 A at 440 V a.c., and 1 A at 240 V d.c. (non-inductive).

Electrical Remote Control Co. Ltd.,
Bush Fair, Tye Green,
Harlow New Town, Essex.



Frequency and Time Measuring Equipment

Venner Electronics have developed a frequency and time measuring equipment, type TSA3336, in which the quantity being measured is displayed on the front panel in digital form.

The TSA 3336 is designed for the measurement of frequency (events per unit time up to 1·1 Mc/s), inverse frequency (period), simple counting in decade steps from 1 to 10^9, and time in decade steps from 1 μsec to 10 sec. Accuracy of the instrument is within ± 1 count and the crystal stability of 1 part in 10^7.

In addition, the outputs from the temperature-controlled megacycle crystal oscillator and the decade divider chain are made available for external use as 'clock' pulses. The instrument can be re-set remotely or locally by manual or automatic operation.

The TSA 3336 includes transistor plug-in circuit modules and can be supplied with outputs for the operation of a printer or re-perforator. It measures 14¼ in. × 12⅞ in. × 8¼ in. high, weighs 26 lb and can be operated from 100—125 V, 200—250 V, 50-c/s mains or 12 V d.c.

Venner Electronics Ltd.,
Kingston By-Pass,
New Malden, Surrey.

A.C. Microvoltmeter

Marconi Instruments' latest A.C. Microvoltmeter Type TF.1375 covers the range 5 μV to 15 V in twelve steps; it is fully transistorized and truly portable, with a self-contained battery supply. The frequency-range is 50 c/s to 1 Mc/s with switch-selected l.f. rejection.

Basically, the instrument consists of a frequency-compensated input attenuator with decade steps, and a seven-transistor frequency-compensated amplifier having a voltage gain of approximately 5,000. Output from this amplifier is rectified and fed to the meter; in addition, the output from the penultimate stage is available at a front-panel outlet.

Input impedance and sensitivity are such that it can be used as a microammeter for detecting currents as low as 10^{-13} A.

Marconi Instruments Ltd.,
St. Albans, Herts.

Wideband Sine-Wave Generator

A compact sine-wave generator covering most audio and low-video applications has been announced by Waveforms Inc., U.S.A.

Known as Model 401F, it covers the frequency range 0·9 c/s to 1·2 Mc/s in 6 decade steps and provides an output of 10 V into a 600-Ω load and 20 V into an open circuit. A 5-position decade attenuator sets the output at 10, 1, 0·1, 0·01 and 0·001 V max. when terminated into 600 Ω, while a continuously-variable fine attenuator covers each range of the coarse attenuator down to $\frac{1}{10}$th maximum output on each decade. The accuracy of the coarse attenuator is ±2% at 1 kc/s and that of the fine attenuator is ±10%. Over the temperature range 0° to 50 °C the frequency stability is ±¼% (9 c/s–120 kc/s), ±2% (120 kc/s–

1·2 Mc/s) and ±5% + 0·5 c/s (0·9 c/s–9 c/s); dial accuracy is ±2% (9 c/s–1·2 Mc/s).

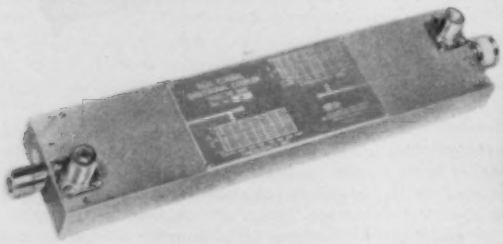
The oscillator operates from 115 V or 230-V mains, it measures 8 in. × 6 in. × 10¾ in., and weighs 12 lb.

Waveforms Inc.,
6 Cornelia Street,
New York 14, N.Y., U.S.A.

Dual Coaxial Couplers

Two Narda microwave dual coaxial couplers covering the range 250 Mc/s to 4,000 Mc/s have been introduced into the U.K. by Aveley Electric.

These coaxial couplers, designated Models 3020 and 3022, have been specially designed for use in coaxial reflectometer assemblies to give a 4 : 1 frequency range while maintaining flat coupling with high directivity. Model 3020 covers the 250–1,000-Mc/s range, while Model 3022 covers 1,000–4,000 Mc/s.



Coupling of the rear arm is held to 20 dB ± 1 dB over the frequency range, and the coupling forward and reverse arms track each other within 3 dB.

A feature claimed for these couplers is an unusually low main and secondary line v.s.w.r. The main line v.s.w.r. in Model 3020 is 1·05 max., while the secondary line is 1·1 max.

Distributed in the U.K. by
Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

Phase-Rotation Indicator

This Elliott phase-rotation indicator has been designed for 3-phase work where it is essential to know before connecting the supply to the apparatus whether or not the sequence of any phase is correct.

The unit is housed in a metal box measuring approximately 1½ in. × 2⅞ in. × 1¼ in. and is fitted with four neon lamps and a short flying lead for connection to the three phases.

If the phase sequence is correct, all four neons will glow when the unit is connected to the supply. If there is any failure of the phase or a reversal in sequence, at least one lamp will fail to glow.

Two models are available for 115-V 400-c/s and 400-V 50-c/s operation.

Elliott Brothers (London) Ltd.,
Airport Works, Rochester, Kent.

Valve Voltmeter

Taylor Electrical have recently announced the release of a mains-operated portable valve voltmeter known as Model 172A.

It has seven d.c.-voltage ranges from 0–1·5 V to 0–1,500 V (30 kV with e.h.t. probe, Model 472), seven a.c. r.m.s.-voltage ranges from 0–1·5 V to 0–1,500 V, seven a.c. peak-to-peak voltage ranges from 0–4 V to 0–4,200 V, and seven resistance ranges from 0–1 kΩ to 0–1,000 MΩ. The input impedance is 11 MΩ on all d.c.-voltage ranges and 830 kΩ to 1·5 MΩ on a.c.-voltage ranges. In conjunction with a small crystal probe the instrument can be used to measure potentials at frequencies above 200 Mc/s.

Additional features of the 172A are 'centre zero setting' to facilitate alignment of f.m.-discriminator circuits and a 'reverse polarity' switch position which reverses the polarity of the meter.

Measuring 7½ in. × 6¼ in. × 4½ in., the instrument weighs 5½ lb. and will operate from a.c.-mains voltages from 110 to 240 V, 40–60 c/s.

Taylor Electrical Instruments Ltd.,
Montrose Avenue, Slough, Bucks.

Ceramic-and-Metal Line Protector

By developing new constructional and sealing techniques, A.E.I. are now able to offer a gas-filled line protector using only ceramic and metal parts.

Designated Type 16, it is of three-electrode design, is smaller than the conventional two- or three-electrode device, yet gives improved performance and, moreover, it can replace two or three two-electrode protectors.

It is available in three ranges of d.c. striking voltages 150–350 V (Type 16A), 300–500 V (Type 16B) and 600–900 V (Type 16C) and it measures 49 × 9 mm. Operating delay is less than 1 μsec when 1,500 V d.c. with a rise-time of less than 0.1 μsec is applied. The protector will withstand without damage ten 1-sec applications of 20 A a.c. every three minutes on any one section. It will also withstand at least 150 mA a.c. for three hours. Insulation resistance measured at 85 V is greater than 1,000 MΩ between any two electrodes; arc voltage is of the order 20 V and on any gap it will withstand many hundreds of 100-joule discharges.

Mountings for the Type 16 protector are available in unit and strip form, and as cable terminals or pot-head insulators.

Associated Electrical Industries Ltd.,
Telephone Apparatus Department,
Woolwich, London, S.E.18.

Surface Demagnetizer

A surface demagnetizer has been introduced by Marinair to demagnetize mild steel to a depth of approximately ½ in., high-speed steel to ¼ in., and magnetically harder materials to a lesser depth.

The unit is encapsulated in epoxy resin and is fitted with 3 yd of 3-core cable; it is available for all a.c. supply voltages, including low-voltage safety systems.

Marinair (Radar) Ltd.,
Eastern Industrial Estate,
Harlow, Essex.

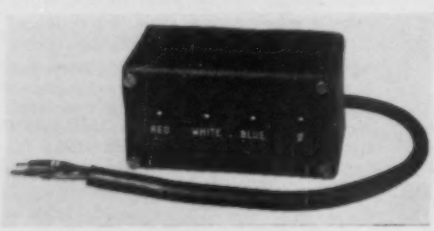
Polyester and Cellulose-Acetate Tapes

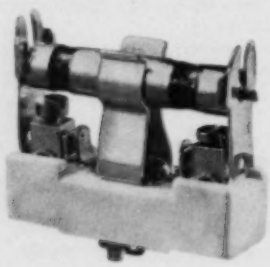
Two additional adhesive insulating tapes for the electrical and allied industries have been introduced by Gosheron.

The first is Transonex polyester thermo-setting tape which is coloured yellow. After curing, the tape will operate at temperatures up to 180 °C and has a breakdown voltage of 5,000 V. Maximum solvent resistance is obtained by curing at 150 °C for 2 hr or 120 °C for 4 hr.

The second is a cellulose-acetate electrical tape which is based on a flexible grade of transparent cellulose-acetate film and has a key-coated electrical-grade adhesive with freedom from corrosive tendencies. These properties make it suitable for use in the manufacture of small coils, where thinness and high-insulation resistance are also required.

John Gosheron & Co. Ltd.,
The Packaging and Industrial Tape Centre,
Albert Embankment,
Vauxhall, London, S.E.1.



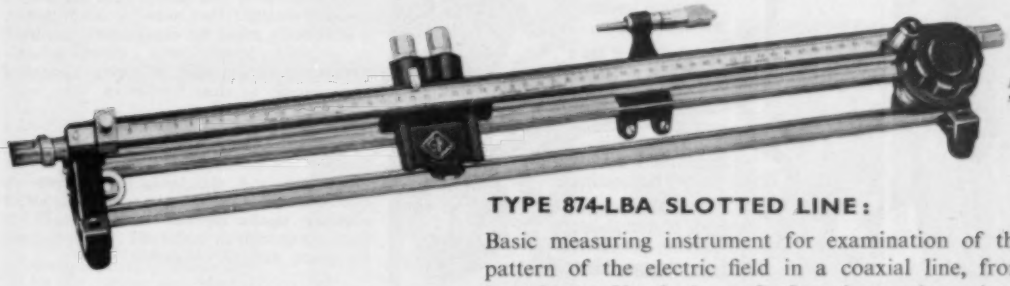


'GENERAL RADIO'

V.H.F. – U.H.F. MEASURING EQUIPMENT

With basic measuring instruments, such as the Slotted Line or Admittance Meter, generators, detectors, and a wide range of inter-related coaxial elements, all linked through the ingenious Type 874 connector, 'GENERAL RADIO' offer the scientist and engineer a 50-ohm VHF-UHF measuring system that is

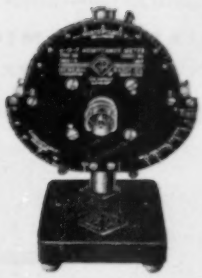
Complete · Integrated · Accurate · Versatile



Micrometer Vernier shown attached.

TYPE 874-LBA SLOTTED LINE:

Basic measuring instrument for examination of the standing-wave-pattern of the electric field in a coaxial line, from which VSWR, impedance of load, phase of reflected wave, losses in attached elements, degree of mis-match between line and load, etc. can be determined. Accurate and straightforward in use. Frequency range 300-5000 Mc/s (with some loss in accuracy: 150-7000 Mc/s). Also available: Type 874-LV Micrometer Vernier (for measurement of high VSWR). Type 874-MD Motor Drive for oscillographic display of standing-wave pattern.



TYPE 1602-B ADMITTANCE METER:

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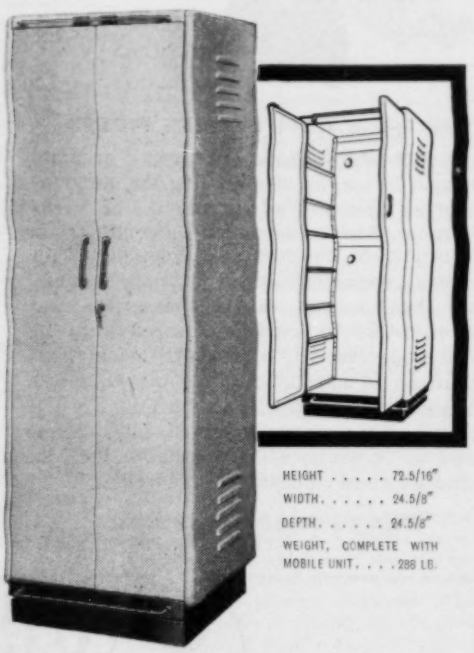

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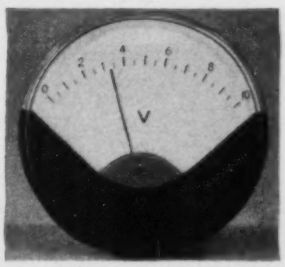

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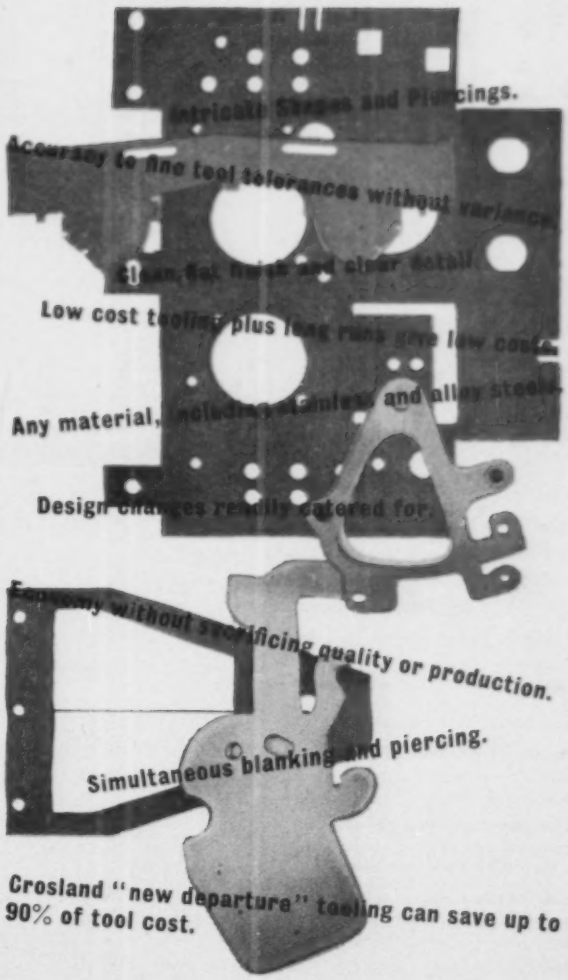
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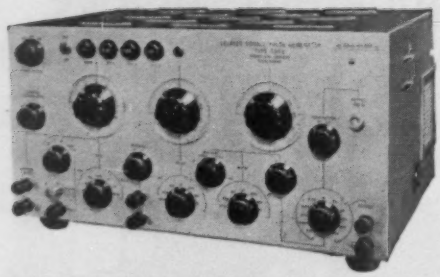

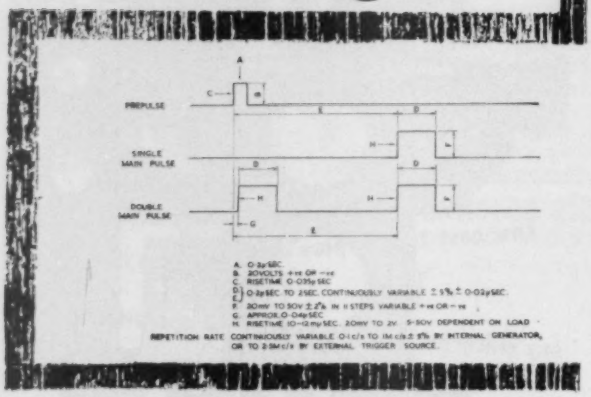

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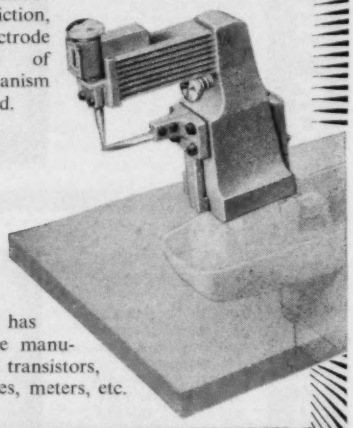

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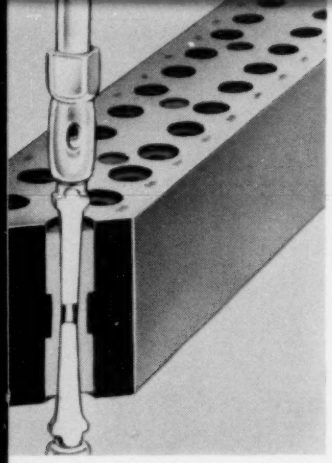

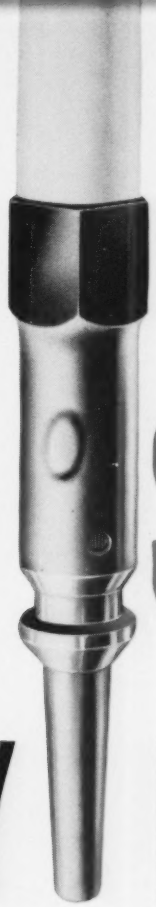

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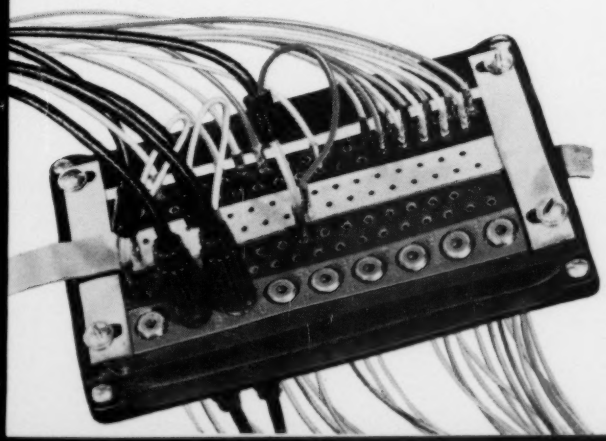

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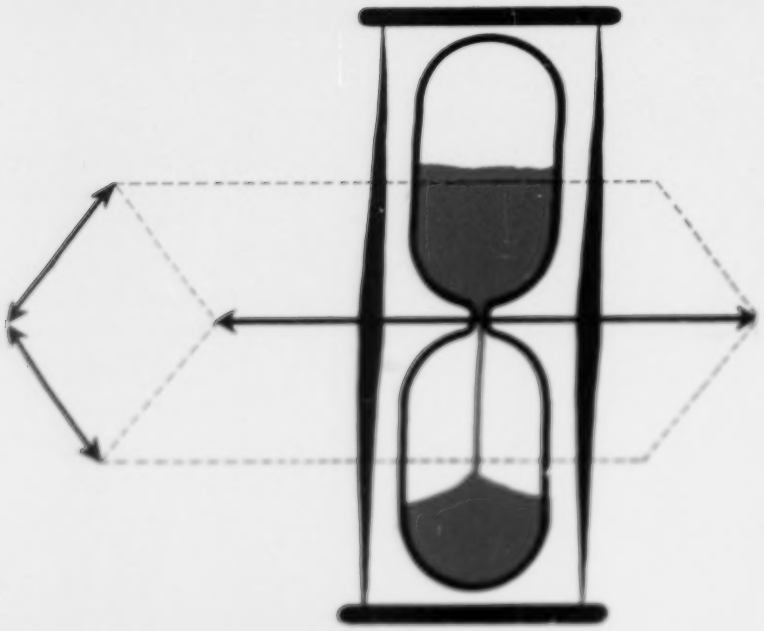

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